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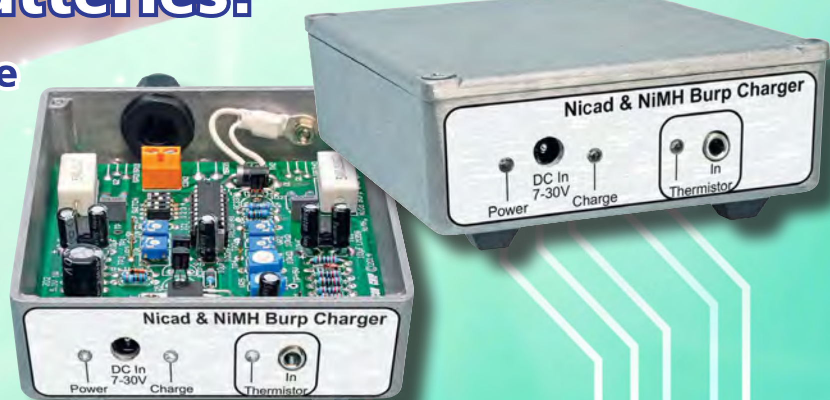

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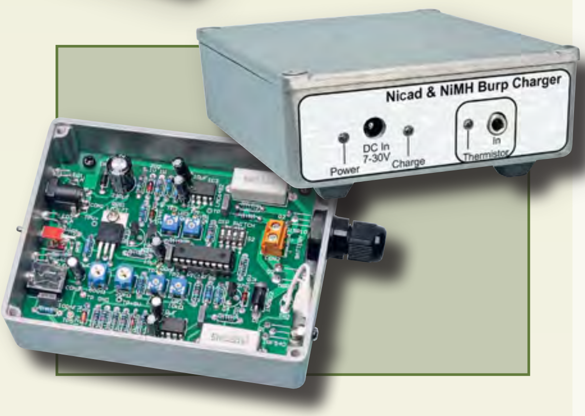

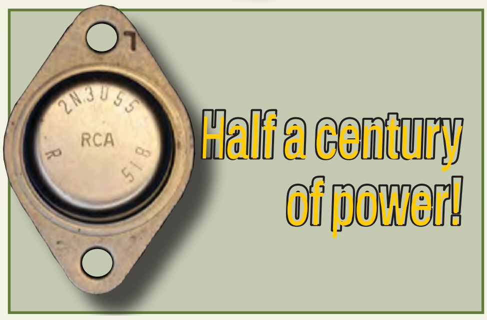

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Our July 2015 issue will be published on Thursday 4 June 2015, see page 72 for details.

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EVERYDAY PRACTICAL ELECTRONICS

The next big thing

We are going to hear a great deal about a couple of little devices in the next few months. First there is Apple's new Watch. Although I must confess I am a serial consumer of Apple products, I won't bother you with much coverage – unless you live under a rock you will be bombarded with opinions and one more piece from me won't make any difference. However, I'll say just this, even if the first iteration of Apple's Watch fails to grab your attention, I would keep an eye on this general technology area. I am in no doubt that it represents one of the next big areas for exciting growth in electronics. If done well, it offers the possibility of real and useful health monitoring to name just one important potential use.

BBC Micro Bit

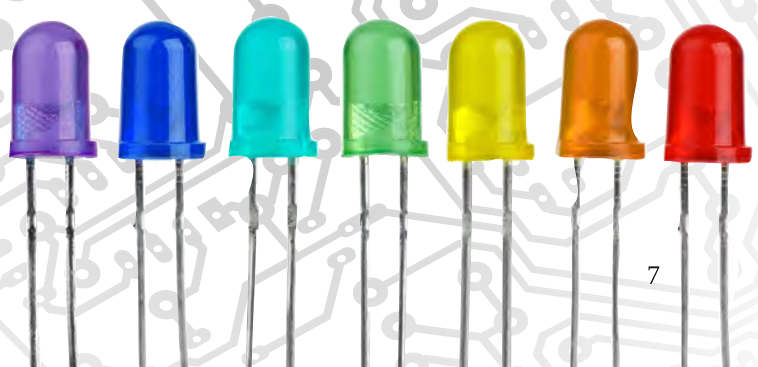
What I would like to flag up again (it was in last month's news) is the BBC's Micro Bit. A tiny, portable and cheap computer that will be given to a million little Britons in school Year 7 in the autumn. It is designed to provide a first step in using a computer and program writing – am I the only one who finds the newly fashionable term 'coding' irritating?

Micro Bit is relatively basic – an IC with a simple LED matrix and a single micro-USB port, all powered by a watch cell. To get it to do something interesting you plug it into a computer – and from there, the idea is that children program the LEDs to flash their initials or a simple message (hopefully, not just 'OMG'). After that, they can move on to more advanced projects involving sensors and a servo to drive an electric motor.

Create/consume

There are two key things I like about this project. First, it helps to reverse the tendency for all of us, but particularly children to be mere consumers of digital technology, and helps them become creators. Second, just like the BBC Micro back in the 1980s, I hope this will eventually be seen as a farsighted investment in the future. Instill the next generation with the thrill of engineering creativity, and I am sure the few million the BBC is spending on this will be repaid a hundred fold in future companies that produce products that are *Made in Britain*.

So, if you have children, grandchildren or perhaps younger siblings who bring a Micro Bit home, then I hope you will help and encourage them to have fun and take their first steps in electronics.



NEWS

A roundup of the latest Everyday News from the world of electronics

Cinema immersive sound – report by Barry Fox

The upcoming launch of DTS:X to rival Dolby Atmos makes the competition to set a standard for immersive surround sound – with height – a three-horse race. The third and darker horse is Auro 3D, a system developed ten years ago. I made a trip to Belgium, where Auro was born, to learn more.

Auro 3D is the brainchild of musician/engineer Wilfried Van Baelen, who – with his brother Guy – started the Galaxy recording studio complex at Mol, deep in Belgian farmland a couple of hours drive from Brussels, see: **www.galaxystudios.com/about-us**

Auro-3D speaker system used for showing the World War II film Red Tails

Studio background

Van Baelen started as a musician playing Hammond organ in the mid-1970s, and built a recording studio in his chicken shed.

There are now 14 interconnecting studios and control rooms, all visually interconnected and with windows for daylight, but each acoustically isolated by 100dB. The rooms are made of concrete and supported on springs, with windows of glass 11cm thick, weighing a tonne.

'When we tried to order the glass' Van Baelen recalls, 'they asked what weapon we were up against. It was after a spate of bank robberies. So we were lucky they had already done some development work.'

Van Baelen started work on Auro ten years ago. 'For an immersive experience you need two layers of speakers, or three layers with 'voice of god' speakers overhead' he says. Auro's coding technology lets a backwards-compatible 5.1 surround mix carry the extra height channels.

Breaking into Hollywood

'We demonstrated Auro to George Lucas in 2011 and he said he would use it to make Red Tails, the first Hollywood movie in immersive sound. The first commercial cinema installation was in Chengdu, China in 2011, followed by Dallas, Los Angeles and Miami in the USA.'

Auro currently claims over 300 installations worldwide. Most are in the US, Asia, India and Europe – but none in the UK. Auro claims 125 movies confirmed for release in Auro 3D, and over 75 released, including Elysium, I, Frankenstein and Spiderman 2.

Auro's business model is a one-off licence for a theatre decoder upgrade sold by partner company Barco (famous for projectors). The first home systems, for the high end from Auriga, Denon, Marantz, McIntosh and Trinnov were rolled out through 2014. Owners pay 160 euros to upgrade them with the Auro engine.

Unlike the old matrix surround systems, which worked in the analogue domain, Auro's system works in the PCM digital domain. The horizontal and height signals are combined in the bit stream, along with coded instructions for the decoder to un-combine them. The 'enhanced 5.1' or Native Auro-3D plays normally on a standard 5.1 system, to give a balanced mix of horizontal 5.1 and height channels, but it decodes with an Auro decoder to give separate Hi Res horizontal and height channels.

Software solutions

Processing latency is one PCM sample per channel – totalling well under one millisecond – so there is no loss of lip sync.

The trick – protected by Auro's original patents – is to use the four least-significant bits of a 24-audio signal to carry the encode/decode instructions.

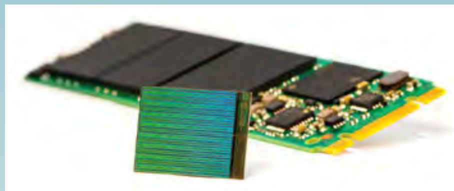
'It's all done in software – you can render channel-based or object-based audio in the same way' explains Van Baelen. Auro currently uses channel-based audio, with up to 15 channels.

Unhealthy rivalry?

There is clearly very strong rivalry between Auro and Atmos. Van Baelen showed me a detailed 171-page Powerpoint presentation prepared for the trade. This repeatedly, and unfavourably, compares Dolby Atmos price and performance with Auro. Needless to say Dolby refutes Auro's conclusions.

When DTS:X finally launches (it was due in March but missed the mark) there seems sure to be a full scale standards battle – which will very likely depress sales for all of the systems.

3D NAND Flash memory



Gum stick-sized SSDs with 3.5TB of storage

Micron Technology and Intel have revealed their 3D NAND technology, claimed to be the world's highest-density Flash memory – the storage technology used inside the lightest laptops, fastest data centers, and nearly every cellphone, tablet and mobile device.

This new 3D NAND technology, which was jointly developed by Intel and Micron, stacks layers of data storage cells vertically with extraordinary precision to create storage devices with three times higher capacity than competing technologies. This enables more storage in a smaller space, bringing significant cost savings, low power usage and high performance to a range of mobile consumer devices as well as the most demanding enterprise deployments.

Planar NAND flash memory is nearing its practical scaling limits, posing significant challenges for the memory industry. 3D NAND technology is poised to make a dramatic impact by keeping Flash storage solutions aligned with Moore's Law, the trajectory for continued performance gains and cost savings, driving more widespread use of Flash storage.

The new 3D NAND technology stacks Flash cells vertically in 32 layers to achieve 256Gb multilevel cell (MLC) and 384Gb triple-level cell (TLC) die that fit within a standard package. These capacities can enable gum-stick-sized SSDs with more than 3.5TB of storage and standard 2.5-inch SSDs with greater than 10TB.

Because capacity is achieved by stacking cells vertically, the individual cell dimensions can be considerably larger. This is expected to increase both performance and endurance and make even the TLC designs well-suited for data center storage.

IBM speed record

IBM has announced a technological leap that will help lift Internet speeds to 200-400 gigabits per second at very low power.

The speed boost is based on a device that can be used to improve transferring 'big data' between clouds and data centers four times faster than current technology. At this speed, 160GB, the equivalent of a two-hour, 4K ultra-high definition movie or 40,000 songs, could be downloaded in only a few seconds.

As big data and Internet traffic continues to grow exponentially, future networking standards have to support higher data rates. For example, in 1992, 100 gigabyte of data was transferred per day, whereas today, traffic has grown to two exabytes per day, a 20-million-fold increase.

Wristwatch Enigma

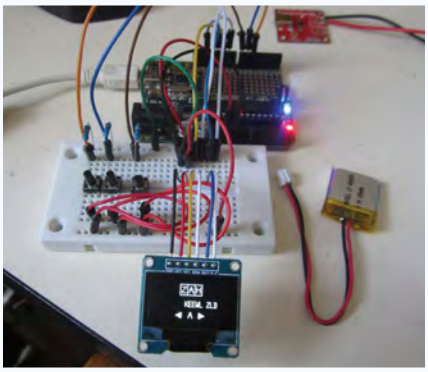
Maker Simon Jansen, a talented New Zealand electronics enthusiast has created a retro-looking three-rotor Enigma wristwatch. Yes, *that* Enigma, the famous German World War II cryptography machine that Alan Turing and colleagues at Bletchley Park cracked.

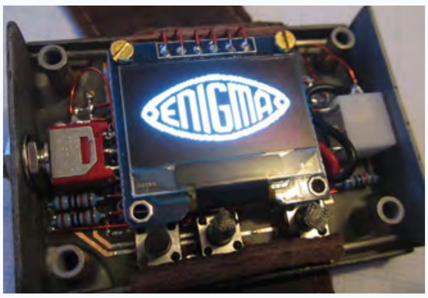
Simon was inspired to build his device after a trip to the UK, which included a visit to Bletchley Park. He was particularly impressed with the Bombe Machine – an electro-mechanical device used by British cryptologists to help them decode Enigma messages. He decided to build his own by coding it in an 8-bit computer he had built. From there, he designed his own Enigma machine using just an Arduino with a small OLED screen, some switches and a LiPo battery.

His aim wasn't to make the smallest Bombe possible, but simply a device that was practical and useable. Something that could actually be used in the field in place of a real Enigma machine. He acknowledges there were some obvious limitations; he couldn't have a 26-key keyboard, so he came up with a user interface that would work with a minimal number of keys.

The cherry on the cake was to make it look as far as possible like it was from WWII. He finished the case in a black crackle paint, used brass push-buttons and even etched an Enigma badge as a finishing touch.

The details are covered at Simon's website: **www.asciimation.co.nz/bb**, with a comprehensive collection of photographs detailing construction.





Silicon kilo

We tend to think of silicon as the ultimate electronics material, but an international team of researchers has developed a new mechanical use – defining the kilogramme, one of the seven fundamental metric SI units (along with the ampere, mole, kelvin, metre, candela and second).

In the 'Avogadro project', the new kilo uses Avogadro's constant and Planck's constant to count atoms in a silicon sphere that is so perfect its shape deviates by less than 100nm – less than a quarter the wavelength of the shortest wavelength of visible light. It has been described as 'the roundest object ever made'.

The kilo's single crystal of silicon from the Electrochemical Plant in Zelenogorsk (Russia) is ultra pure and has been refined to 99.998% purity.

Churchill's Scientists

If you find yourself near South Kensington in London and have an hour or two to spare, then 'Churchill's Scientists', a free exhibition at the Science Museum is well worth a visit. It is a fascinating look at the contribution British scientists and engineers made to the war effort – including, of course, in electronics.

Touch-Screen Digital Audio Recorder – Part 1

By ANDREW LEVIDO

Want to record and play back with CD sound quality using a compact hand-held unit with a colour touch-screen? Well, now you can. This device records to and plays back from a standard SD card and doubles as an SD card reader when connected to your PC via its USB interface. A single AA-size lithium-ion cell provides hours of record or playback time and is recharged via the USB port.

WHAT'S THE FIRST thing you will notice about this *Digital Sound Recorder*? It has no external controls! Just like smart phones and tablets, everything is done via the touch-screen. All its inputs and outputs are at the top end of the case – stereo line inputs with adjustable gain, a mono external microphone input jack and an in-built electret microphone, with two settings for gain (again, via the Touch-screen).

Audio output is via a stereo line output jack (3.5mm socket) and a headphone jack (3.5mm socket) with its volume adjustable via the touch-screen. Also at the end of the case is the SD card socket and a single LED that indicates when card read or write operations are in progress. And there is a mini-USB socket for communicating with a PC and charging the battery.

It records and plays standard WAV-format audio files and is compatible with any PC or Mac. It supports 16-bit stereo PCM-coded files at sample rates from 8-96ks/s.

The touch-screen display is a 72mm (diagonal) QVGA (quarter VGA or 320 × 240 pixels) TFT model. It has a white LED backlight and supports 262,000 colours (although only 65,000 are allowed for by the software).

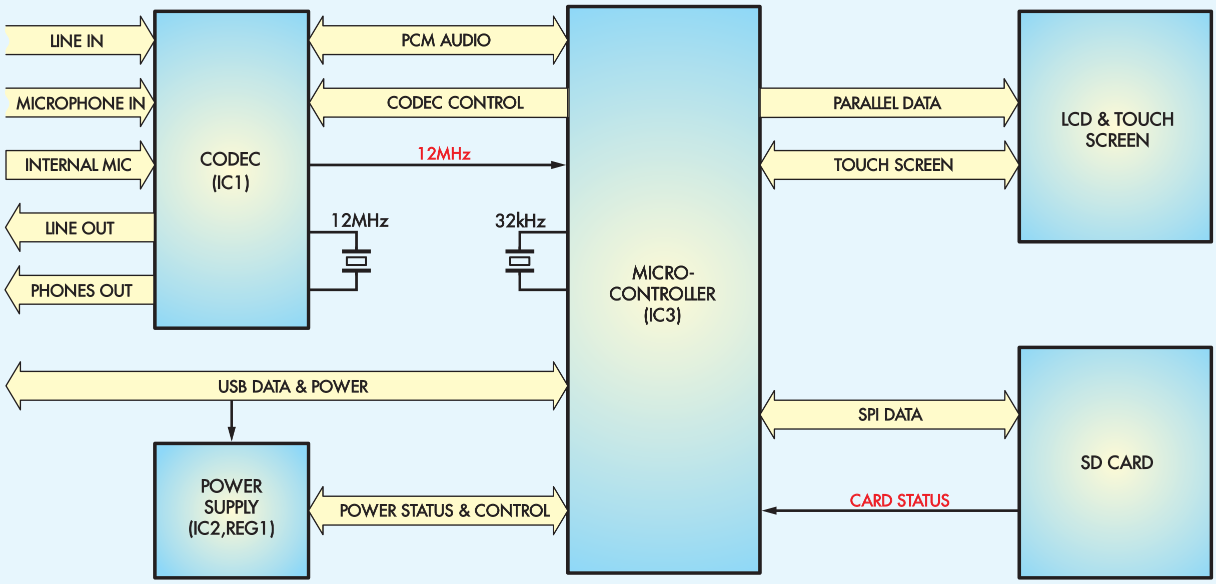


Fig.1: the block diagram of the *Touch-Screen Recorder*. A TLV320AIC23 CODEC (IC1) takes care of all the analogue signal processing, plus analogue-to-digital and digital-to-analogue conversion of the audio streams. This interfaces to a PIC32 micro (IC3) via two serial data paths (PCM audio and CODEC control) and the micro in turn drives an LCD touch-screen display and an SD card. Microcontroller IC3 also provides USB support.

We have made the user interface intuitive, with on-screen buttons and text, and the display also shows the date, time and battery-charge state.

To conserve the lithium cell, the backlight automatically dims after 30 seconds of touch-screen inactivity, and it immediately brightens again when the screen is touched. The recorder goes to sleep after a further 30 seconds of touch-screen inactivity, provided it is not recording or playing and is not connected to a USB power source. Simply touching the screen wakes it up again.

When a PC is connected, the recorder can be put into SD-card-reader mode. The SD card will appear to the PC (or Mac) as an external disk drive, so files can be transferred back and forth. Mind you, to transfer a lot of files it will be quicker to remove the SD card from the recorder and insert it directly into your computer or a dedicated card reader.

So why would you bother to use this recorder rather than using your smartphone? The quick answer is great sound quality. This recorder gives you CD sound quality which your smartphone simply cannot!

How it works

For all of its fancy features, the *Touch-Screen Recorder* only uses a couple of chips. Basically, all it has is a CODEC (coder-decoder), a PIC32 microcontroller and the LCD touchscreen This is shown in the block diagram of Fig.1. The TLV320AIC23 CODEC takes care of all the analogue signal processing, plus analogue-to-digital and digital-to-analogue conversion of the audio streams.

This interfaces to the PIC32 microcontroller over two serial data paths, one bidirectional path for the PCM audio data and one single-direction path for CODEC control. Two of the microcontroller's three Serial Peripheral Interface (SPI) modules are used for these interfaces. The CODEC requires a 12MHz crystal for timing and provides a buffered clock output, so we have used this to provide the main clock input for the microcontroller.

The microcontroller's third SPI module is for communication with the SD card. Two digital inputs monitor the state of the card presence and write protect switches in the SD card socket.

The LCD is driven via an 8-bit parallel interface with read, write and chip select lines. Although mechanically integrated with the display, the touchscreen is electrically separate and is essentially an analogue device, so it is connected to pins on the micro that can double as analogue inputs.

More details of how the touchscreen works are given later.

Main Features

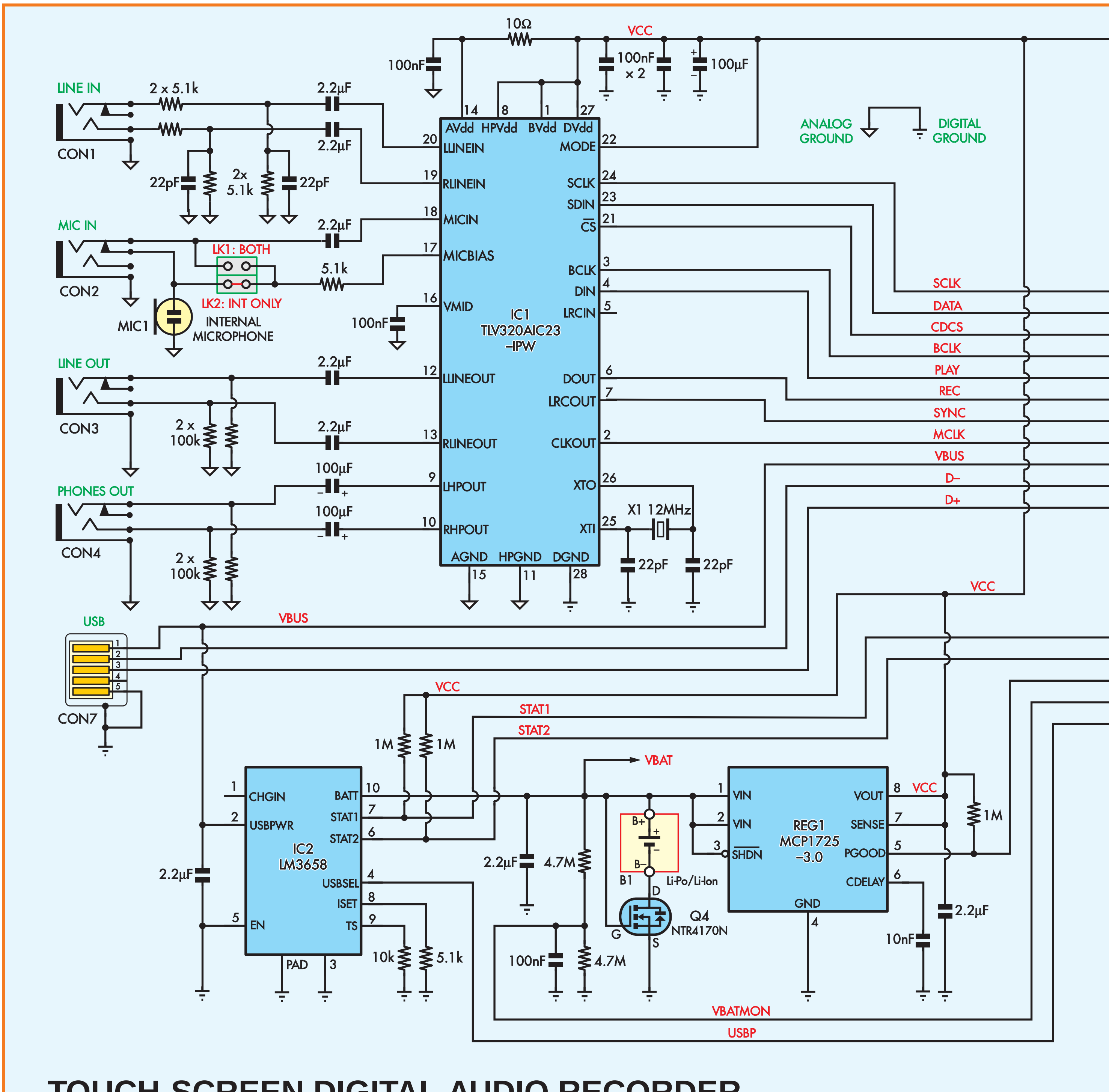
- CD sound quality
- Colour touch-screen with no external controls
- SD card memory
- Powered by a single AA-size lithium-ion cell; recharged via an on-board USB port

The USB socket connects directly to the microcontroller and the 5V USB bus power feeds a dedicated lithium-ion battery charger IC. The battery voltage can vary from 4.2V to 3.2V and is regulated to provide a 3V rail for all of the electronics (except for the display backlight that uses the unregulated battery supply).

The microcontroller monitors the battery voltage via an ADC input and this, together with status outputs from the battery charger and regulator, allows the micro to display battery status.

More detail

Let's now refer to the main circuit diagram for more details – see Fig.2. At the top lefthand corner, we can see that the stereo line input jack (CON1) is connected to the CODEC line input, pins 19 and 20, via voltage dividers



TOUCH-SCREEN DIGITAL AUDIO RECORDER

and 2.2µF blocking capacitors. These dividers ensure that the input impedance is approximately 10kΩ and attenuate the line level signal, which can be as high as 2V RMS, to a maximum of 1V RMS – the full-scale input level for the CODEC.

The CODEC contains a digital gain/attenuation stage for the line input that can be set to any value between –34.5dB and +12dB in 1.5dB steps. The line inputs can be muted under control of the micro (IC3).

The mono microphone input (CON2) is connected to the CODEC via another 2.2µF DC blocking capacitor. The on-board electret microphone element is connected to the switch terminal on the microphone jack so that it is switched out of circuit when an external microphone is plugged in. Pin 17 of the CODEC provides a low-noise DC output to bias an electret microphone.

This can be connected either via link LK1, labelled BOTH, to both the internal and external microphones or via Link LK2 (labelled INTL) to just the internal microphone.

The CODEC provides a fixed +14dB microphone gain stage, followed by an additional +20dB gain stage that can be switched in or out under software

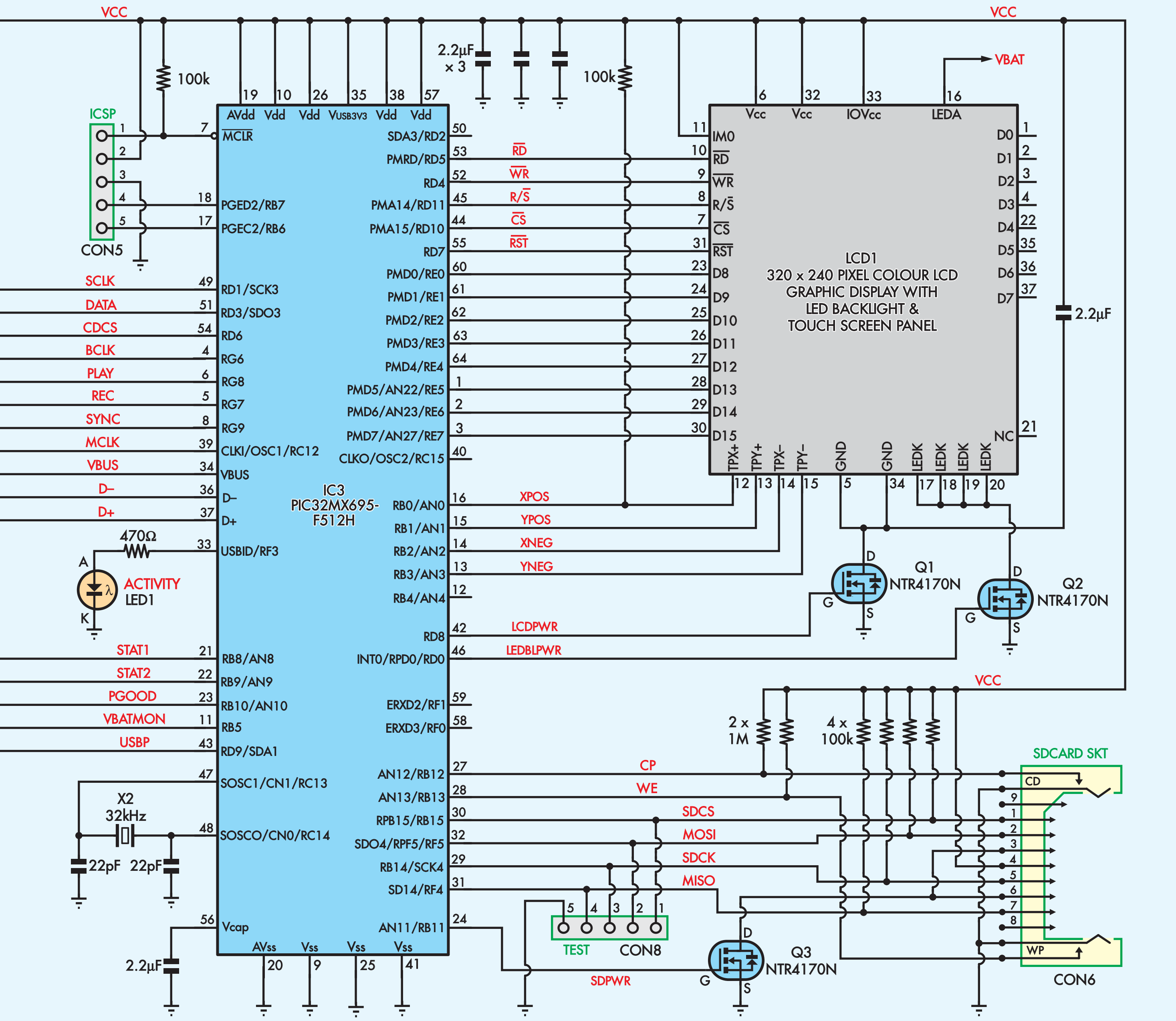
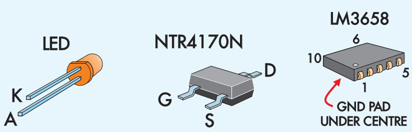


Fig.2: the complete circuit diagram for the *Touch-Screen Recorder.* It's based on CODEC IC1, PIC micro IC3, touch-screen display LCD1 and an SD card. IC2 (LM3658) provides the charge current to the lithium-ion cell (when the device is connected to a USB port) that's used to power the device. The recorded audio data is stored on the SD card and played back under the control of IC3.



control (ie, via the touch-screen). The microphone input can also be muted under software control.

For the audio outputs, the CODEC's internal DACs feed line output buffers that provide fixed-level line outputs on pins 12 and 13. The 2.2µF blocking capacitors prevent any DC bias appearing at the line output jack (CON3), and 100kΩ resistors ensure that the outputs remain referenced to ground. The DAC outputs are also fed to an internal headphone amplifier whose gain is digitally adjustable from –73dB to +6dB in 1dB steps. 100µF blocking capacitors ensure decent low-frequency response, with low-impedance headphones.

If required, the CODEC's audio inputs can be routed to its outputs to provide an analogue bypass path. We use this feature to allow monitoring of the signal being recorded. During playback, this bypass function is switched off.

All of the audio circuitry within the CODEC is referenced to a voltage mid-way between the positive supply and ground. This voltage appears at the VMID pin of the CODEC (pin 16) and is bypassed by a 100nF capacitor.

The CODEC has four different power supply inputs, two for the digital parts of the circuit and two for the analogue

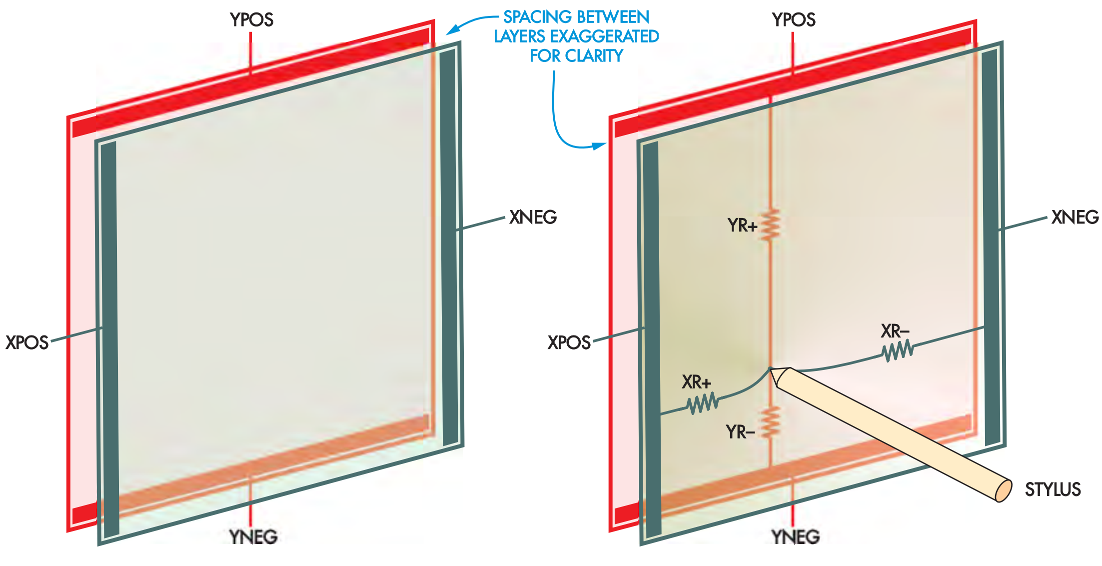


Fig.3: the resistive touch screen is formed using two separated layers of plastic film, each coated with a transparent resistive compound. These effectively form two wide resistors, one in the horizontal (X) direction and one in the vertical (Y) direction. When the screen is pressed, the two films touch and this creates a pair of voltage dividers joined at the point of contact.

parts. The two digital supplies, DVdd (pin 27) and BVdd (pin 1), are connected directly to the 3V rail, as is HPVdd (pin 8), the supply for the headphone amplifier. These are bypassed by a pair of 100nF ceramic capacitors and a 100µF electrolytic capacitor.

The supply for the analogue circuitry, AVdd (pin 14), is derived from the 3V rail (VCC) via a simple RC filter consisting of a 10Ω resistor and a 100nF capacitor. Care has been taken with the PCB layout to ensure that power supplies and ground planes for the analogue and digital parts of the circuit are connected so as to minimise the conduction of digital noise into the sensitive analogue circuitry.

The digital side of the CODEC consists of a standard digital audio interface on pins 3-7. In our case, the CODEC is configured to be the master. It outputs a bit clock on pin 3 and a frame sync pulse on pin 7.

The frame sync pulse indicates the start of a data frame that consists of the left and then right data words. The bit clock operates at 12MHz or 6MHz, depending on the data rate selected, and the frame rate is equal to the sample rate. For example, at 48ks/s the bit clock rate is 12MHz and the frame rate is 48kHz.

The record data stream from the CODEC appears on pin 6 and the playback data stream from the micro appears on pin 4. Although the CODEC is capable of a number of different word lengths, we use 16-bit words exclusively. If configured correctly, the SPI module in the microcontroller can operate in a framed slave data mode compatible with this data stream – the challenge is to keep the data flowing fast enough so that the audio stream is played or recorded seamlessly.

As mentioned above, a second SPI port is used to configure and control the CODEC. This is a one-way interface on pins 21-24. Connecting pin 22 to VCC selects the SPI mode (the CODEC also supports I^{2}C for this interface) and the usual chip select, clock and serial data lines are used to receive command data from the microcontroller.

Finally, the CODEC clock is derived from a 12MHz crystal connected to pins 25 and 26, with associated 22pF loading capacitors. The 12MHz clock output from pin 2 drives the microcontroller's clock input to provide the system clock when the microcontroller is awake.

Touch-screen display

As mentioned above, the display is a QVGA colour TFT (thin-film transistor) LCD with white LED backlighting. The display incorporates an ILI9341 driver chip configured for an 8-bit or 16-bit parallel interface. This display-driver chip contains a large number of control registers and a display memory which has one 18-bit data word for each of the 76,800 pixels of the display.

Each 18-bit word defines the 6-bit intensity of each of the red, green and blue sub-pixels, permitting 262,144 colours per pixel. 18 bits is an awkward size to work with given an 8-bit bus, so the display controller allows for the 18 bits to be mapped to a 16-bit word, where each pixel is represented by 5 bits of red data, 6 bits of green data and 5 bits of blue data. This is known as 5:6:5 RGB and permits 65,536 colours which is more than sufficient for our application. As we use an 8-bit interface, it takes two write operations to set one pixel of the display.

The display driver connects to the microcontroller through the above-mentioned 8-bit data bus, a chip select line, and read and write strobes. A single address line on pin 8 of the display determines whether data is written to or read from the control registers or the display RAM. A reset line allows the microcontroller to reset the display to a known state prior to configuring it.

The backlight is driven by a PWM signal from the microcontroller via MOSFET Q2. As noted earlier, the back-light is fed from the unregulated battery voltage, to maximise the possible brightness and to minimise the power dissipation in the regulator.

MOSFET Q1 is used to disconnect the power entirely from the display in sleep mode. We found this necessary since the sleep current of the display was several tens of micro amps.

Touch-screen operation

The touch-screen is physically integrated with the display but is electrically quite separate. This one is a resistive touch-screen and is formed from two layers of plastic film, each coated with a transparent resistive compound and separated by a small air gap. One film has conductive bars printed across the top and bottom edges, while the other has conductive bars printed down each side.

The two plastic films effectively form two wide resistors, one oriented

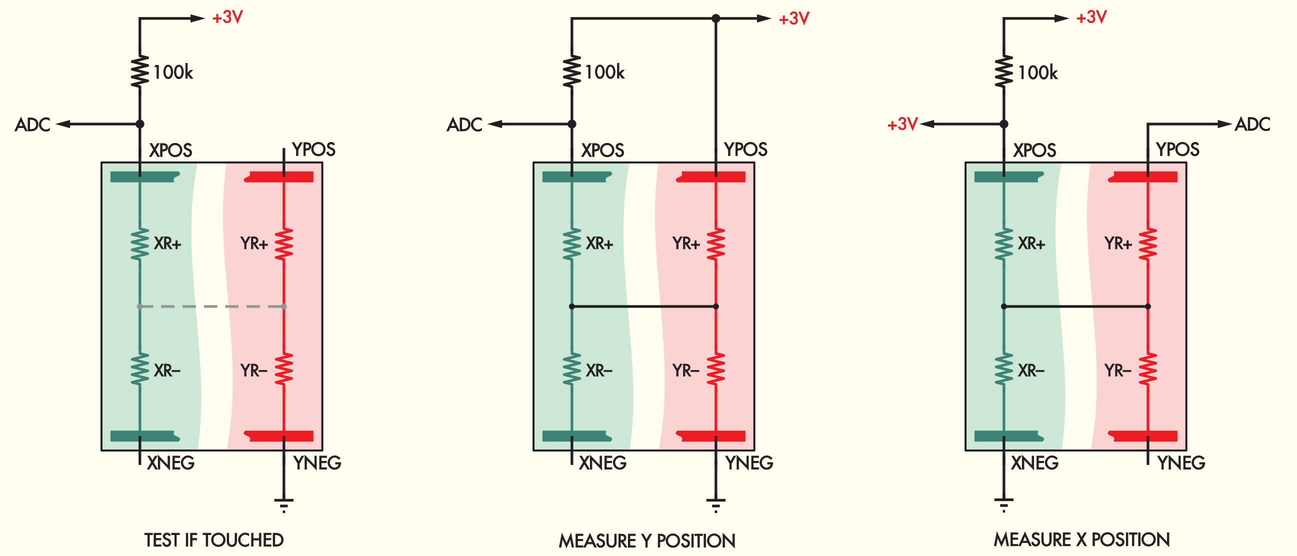


Fig.4: the software polls the touch-screen 100 times per second with the configuration shown at left. If the screen is touched, the ADC will read a low value since the maximum resistance of the touch-screen is much lower than the 100kΩ pull-up resistor. The micro then reads the X and Y positions as shown at centre and right.

in the horizontal (X) direction, and one in the vertical (Y) direction, as shown in Fig.3.

When the screen is pressed with a finger or stylus, the two films touch at the point of contact. This effectively creates a pair of voltage dividers joined at this point. If a voltage is applied between the two X terminals, the voltage measured at one of the Y terminals (while the other is open circuit) will be proportional to the X-coordinate of the touch point. Similarly, if a voltage is applied between the two Y terminals, the voltage measured on one X terminal will be proportional to the Y-coordinate of the touch point.

A bit of scaling and offsetting is required in software. However, it is relatively straightforward to calculate the position of the touch point in terms of the X and Y coordinates of the display pixel on which it occurs. Fig.4 shows how this is done.

Note the 100kΩ pull-up resistor on the XPOS line from the touch-screen. This resistor helps detect when the screen has been touched.

In normal operation, the software polls the touch-screen 100 times per second with the configuration shown in Fig.4 on the left. If there is no touch, the ADC will read a high value. If the screen is touched, the ADC will read a low value, since the maximum resistance of the touch-screen is much lower than the 100kΩ pull-up. If this test shows that the screen is touched, the micro commences the process of reading the X and Y positions, as shown in the centre and right of Fig.4.

When the *Touch-Screen Recorder* is in sleep mode, the ADC is shut down to save power, so we need another way to detect a touch and thus wake the microcontroller. The same configuration of inputs is used as for the touch detection described above, but this time the XPOS input pin is configured to provide an interrupt when there is a change of state. Touching the screen changes the normally high level on this input to logic low, triggering an interrupt which wakes the microcontroller.

SD card interface

The SD card socket (CON6) is connected to the third SPI port on the microcontroller, on the righthand side of the circuit. 100kΩ pull-up resistors are used on each of the lines required by the SD card standard, since some cards apparently power up in open-collector mode. Once the SD card is configured, the outputs switch to totem-pole drivers for speed.

The card socket also has switches for detecting the presence of a card and the position of the write-protect slider. These are also connected to the microcontroller via 1MΩ pull-up resistors.

The SD card's ground pins are switched to ground via MOSFET Q3, which is turned on in normal operation. This allows the SD card to be disconnected in sleep mode, since some cards draw as much as 10mA when idle. MOSFET Q3 is also used to 'hard reset' the SD card if necessary.

PIC32 microcontroller

The other connections to the microcontroller include the in-circuit programming header (CON5), the 32kHz crystal and associated loading capacitors, the USB data and bus voltage sensing lines from the USB socket (CON7) and a single LED to indicate SD card read or write activity. Other pins are used for control and monitoring of the power supply, as described below.

Power supply

The entire circuit, with the exception of the LCD backlight, operates from a 3.0V rail (VCC) derived from the single lithium-ion cell via a low drop-out linear regulator (REG1). This is a Microchip MCP1725 device that can maintain regulation with a drop-out voltage of only 50mV at light load. This is important because the lithium ion cell has a nominal voltage of 3.7V. Fully charged, it produces 4.2V, but as it discharges, its output drops to 3.2V or below.

The regulator has an open-collector output (PGOOD, pin 5) that pulls low if the regulator output falls below 96% of the nominal 3.0V (at approximately 2.88V). The microcontroller software responds to this by ending any recording or playback in progress and putting the *Recorder* into sleep mode. This is necessary to prevent permanent damage to the lithium ion cell, which can occur if it is discharged below 2.7V.

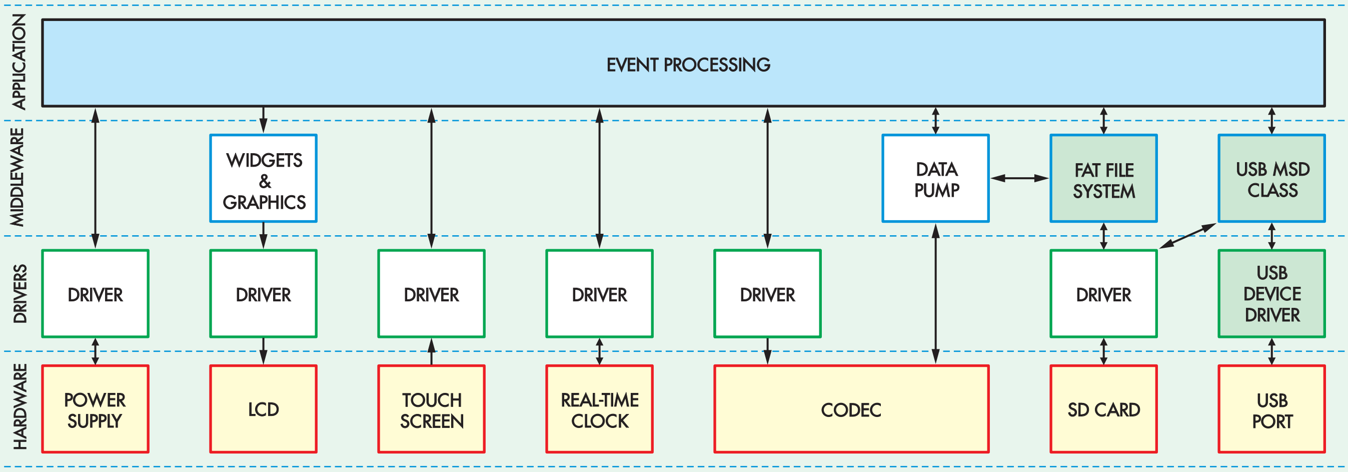


Fig.5: a simplified view of the firmware architecture which is effectively split into three horizontal layers. At the bottom, working directly with the internal peripherals and the external hardware is the 'Driver' layer. This is then followed by a 'Middleware' layer and then an 'Event Processing' layer – see text.

MOSFET Q4 provides polarity protection, against inserting the lithium cell backwards. The cell has a fairly low impedance and could damage the regulator and other semiconductors if inserted wrongly (as we discovered the hard way). A series diode can't be used in this case because the battery has to charge as well as discharge, and in any case, we can't really tolerate the relatively high voltage drop of a diode in this circuit.

The MOSFET is ideal for the job because the channel can conduct in either direction if the gate is positive with respect to the source pin. If the cell is inserted backwards, the MOSFET will be off and the body diode will be reverse biased, so no current will flow.

The microcontroller reads the battery level via a voltage divider consisting of two 4.7MΩ resistors. These high values were chosen to limit the current drain of the divider, which is connected across the battery.

The cell itself is charged by IC2, an LM3658 lithium-ion or lithium-polymer charger, designed to run from USB. It uses a complex, multi-stage charging process and can monitor battery temperature, although we don't use this feature.

IC2 also limits the current drawn from the USB port to 100mA or 500mA, depending on the level of the signal on pin 4, to ensure the current drain remains within USB requirements. USB devices are not supposed to draw more than 100mA until they indicate this requirement and are enumerated by the host.

Pins 6 and 7 of IC2 are open-collector status outputs that indicate the charger state. The software reads these inputs and uses the information together with the cell voltage to display the battery status. When not connected to a USB port or charger, the battery indicator displays HIGH, FAIR or LOW to indicate remaining battery capacity.

When the unit is connected to a charging source, the indicator displays CHRG or FULL or ERR to indicate whether charging is in progress, complete or has failed for some reason. The battery charger indicates an error if the battery cannot be charged within a 5-hour period, usually indicating a faulty battery. If this error occurs, the USB power must be removed and restored to reset the battery charger.

Firmware description

The firmware for the *Touch-Screen Recorder* is relatively complex and a full explanation of its workings is beyond the scope of this article. However, let's provide a broad overview.

Fig.5 shows a simplified picture of the firmware architecture, which is split into three horizontal layers. The green shaded boxes indicate pieces of third-party code that were incorporated into the design. At the bottom level, working directly with the internal peripherals and the external hardware is a series of drivers. The aim of these drivers is to hide the device-specific details and provide a programming interface independent of the hardware. Ideally, we could replace the hardware (say by using a different display) and only have to modify the relevant display driver. Similarly, the same display could be used in another project and the driver should be usable without modification.

By way of example, the LCD driver hides the device-specific instructions necessary to configure the display driver chip behind a 'C' function that initialises the display. Another function draws a single pixel of a specified colour at a point defined by given X and Y co-ordinates. Other functions are available to draw solid blocks of pixels, to shut down and wake up the display and control the backlight. These functions are exposed to the upper layers in the relevant 'C' header files. All the device-specific complexity and any local variables or functions are hidden away within the driver.

Unlike the LCD, some of the drivers have to respond to real-time events (like a touch on the screen). These drivers need some mechanism to let the system know that an event has happened, and the system needs a mechanism to manage and make sense of the unpredictable influx of events.

There are plenty of possible approaches, but we have elected to handle this with an event queue. The drivers post events to a first-in, first-out (FIFO) queue as they occur. They are dealt with in turn by the event processor, which we discuss below.

Middleware layer

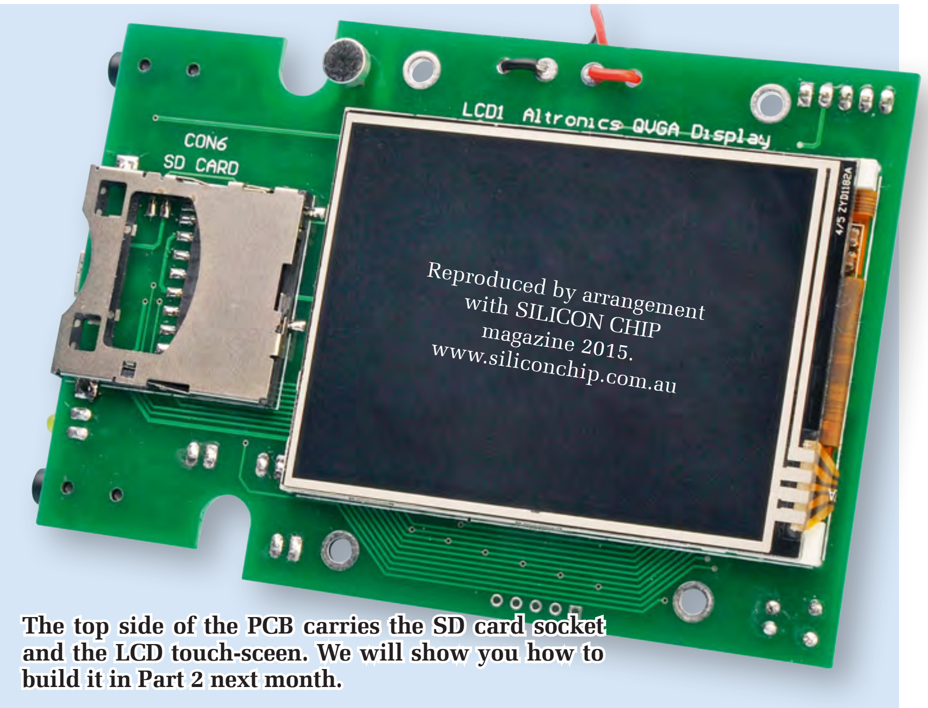
A middleware layer is used when an advanced level of abstraction is required between the drivers and the application. Continuing our example from above, it would be handy to have

a mechanism to draw more complex items on the display, such as lines, shapes, text and icons. These requirements are neither application specific, like the top layer, nor are they hardware specific, like the drivers.

The graphics middleware module, for example, calls on the driver functions that draw a single pixel or block of colour and exposes useful higher-level functions. One such function draws a line between any two points in a specified colour or thickness. Another draws proportional width text.

In fact, this module provides routines for drawing filled and empty circles or arcs, rectangles and round-cornered rectangles, for rendering text and drawing icons. Not all of these are used in the *Touch-Screen Recorder*, since this module was developed for and has been used in other projects.

Also in the middleware layer is the data pump (responsible for shifting data between the CODEC and the SD card), the FAT file system and part of the USB stack. Lets look at these in a bit more detail.



The top side of the PCB carries the SD card socket and the LCD touch-sceen. We will show you how to build it in Part 2 next month.

Data pump and file system

In many ways, the data pump is the beating heart of the *Touch-Screen Recorder*. It has to move data between the CODEC and the SD card at a sufficient rate to avoid glitches in the audio. The CODEC produces (and consumes) data at a rate proportional to the sampling rate, number of channels and the word size. We use a fixed 16-bit data width and two channels, so each audio sample is four bytes long. At 96ks/s we have a data stream of 384,000 bytes per second to contend with.

In contrast, reading and writing files to and from the SD card is a discontinuous process. Data is stored in clusters of 512-byte sectors that may be distributed around the disk in a non-contiguous manner. By the way, the SD card must be formatted with a FAT file system (this is how most cards are formatted out-of-the-box). FAT32, FAT16 and even the older FAT12 formats are supported. Files can be played from or recorded in any directory and file names up to 128 characters are supported.

Each recording is made in a new file that is given a unique name based on the date and time. For example, a recording made on 16 May 2014 at 2:57:20pm would be written to a file named REC140516-145720.WAV.

The file system has to consult the file allocation table on the disk to know where to find or place the next cluster of the file being read or written. All of this takes time. Add to this the fact that SD cards are based on Flash memory that has a relatively long write time.

When we write to the SD card, the data is actually written into an internal RAM buffer (which is fast) but when this buffer is full, the card must write the contents of this buffer to the Flash, a process that can take a few hundred milliseconds or more. The precise time will depend on the write speed of the card and the size of the internal buffer. In general, newer cards are faster than older ones.

Thus, we need to use a pair of internal buffers in the data pump so that one can be filling (to use recording as an example) with data from the CODEC while the other is being written to the SD card. As long as the write time does not exceed the buffer fill time, the system will be ready to write the next buffer as soon as it is filled. If the write time does exceed the fill time, we will find ourselves trying to write and fill the same buffer. This would lead to audible glitches.

The PIC32 microcontroller we have chosen has 128kB of RAM. We found we could allocate up to 96kB of this for the data buffers (two buffers of 48kB each). At the fastest data rate, we fill one of these buffers every 125ms. This is enough to cater to most SD cards. At lower data rates, the SD card write times become less of an issue. At 8ks/s, the buffers each take 1.5s to fill.

The data is shifted between the CODEC SPI and the buffers by DMA (direct memory access), so no processor intervention is required in this process. When one buffer is filled or emptied, the DMA unit automatically switches over to fill or empty the other one and an interrupt is generated. Code in the interrupt service routine takes care of reading or writing the data to or from the SD card. The activity LED (LED1) is lit when this is in progress.

The file system used in this project is called FatFS and was developed by a Japanese hobbyist who goes by the online name of ChaN. This free file system is available at **http://elm-chan.org/fsw/ff/00index_e.html** and has an open license for hobbyist or commercial applications. It has proven to be far more robust, better documented and much faster than other file systems we tried, including one from Microchip.

USB stack

We do, however, use Microchip's USB stack (available free from their website). This consists of a large number of files containing the driver code to control the USB interface peripheral and the higher-level elements of the USB stack necessary to implement a USB 'Device'. In USB-speak, entities can be either a 'Host' (typically a PC) or a 'Device', like

The end panel of the *Touch-Screen Recorder* provides access to the line input and output sockets, the microphone and headphone sockets, the USB socket and the SD card socket.

the *Touch-Screen Recorder.* When first connected, a USB Device makes itself and its capabilities known to the Host through a process called enumeration.

During enumeration, the Device must tell the Host what class of device it is, so that the Host can load the appropriate driver. There are a number of standard Device classes for which common operating systems have native drivers. Examples include, the Human Interface Device (HID) class for computer mice and keyboards; and the Mass Storage Device (MSD) class for hard disks, memory sticks and the like.

The *Touch-Screen Recorder* is configured to appear as an MSD class Device and so will appear to a Windows, Mac or Linux operating system as an external disk drive.

Interestingly, the USB Mass Storage class does not rely on the Device's file system, but rather presents the disk drive to the Host as a SCSI disk and uses the intelligence at the Host end to make sense of the data. The MSD firmware does, however, require the user to provide drivers to handle the basic communication with the media, such as reading or writing a sector.

Fortunately, the requirements of the FAT file system and the MSD system are similar, so only one set of drivers is required. Unfortunately, the MSD stack only ever reads and writes a single sector at a time, whereas the FAT file system makes use of multi-sector reads and writes that are much faster for the bulk transfer of data. This means that data transfer to and from the SD card over the USB interface is relatively slow.

Event-driven programming

At the top level of the architecture, an event manager controls the specific functionality of the *Touch-Screen Recorder* application. The various drivers and middleware modules post messages to the event queue to signal that some particular event has occurred. The event message indicates the source of the event and any important details. For example, the touch-screen driver posts a message if the screen is touched and includes the X and Y coordinates of the point where it occurred.

In the *Touch-Screen Recorder*, events are posted by the touch-screen driver (Press, Still Press, Release, Invalid), the SD card driver (SD Inserted, SD Removed), the real-time clock (One Second Tick, Half Second Tick), the data pump (Recording Stopped, Playing Stopped) and the USB driver (Device Disconnected, Device Detached, Device Attached, Device Ready). Many different error events can also be posted, although these are handled in a slightly different way as described below.

After initialising the various subsystems, the main flow of execution enters a loop where it constantly monitors the event queue. Whenever an event is found in the queue, it is popped out and processed by calling the appropriate routines to handle that event type.

Error events are not posted to the queue in the same way as other events, since event processing is serial and there may several events in the queue ahead of the error event. We don't want to wait until all the pending events are handled before dealing with the error. Using the 'C' standard functions for 'non-local jumps' (<setjmp.h>), we can ensure errors are processed immediately they occur.

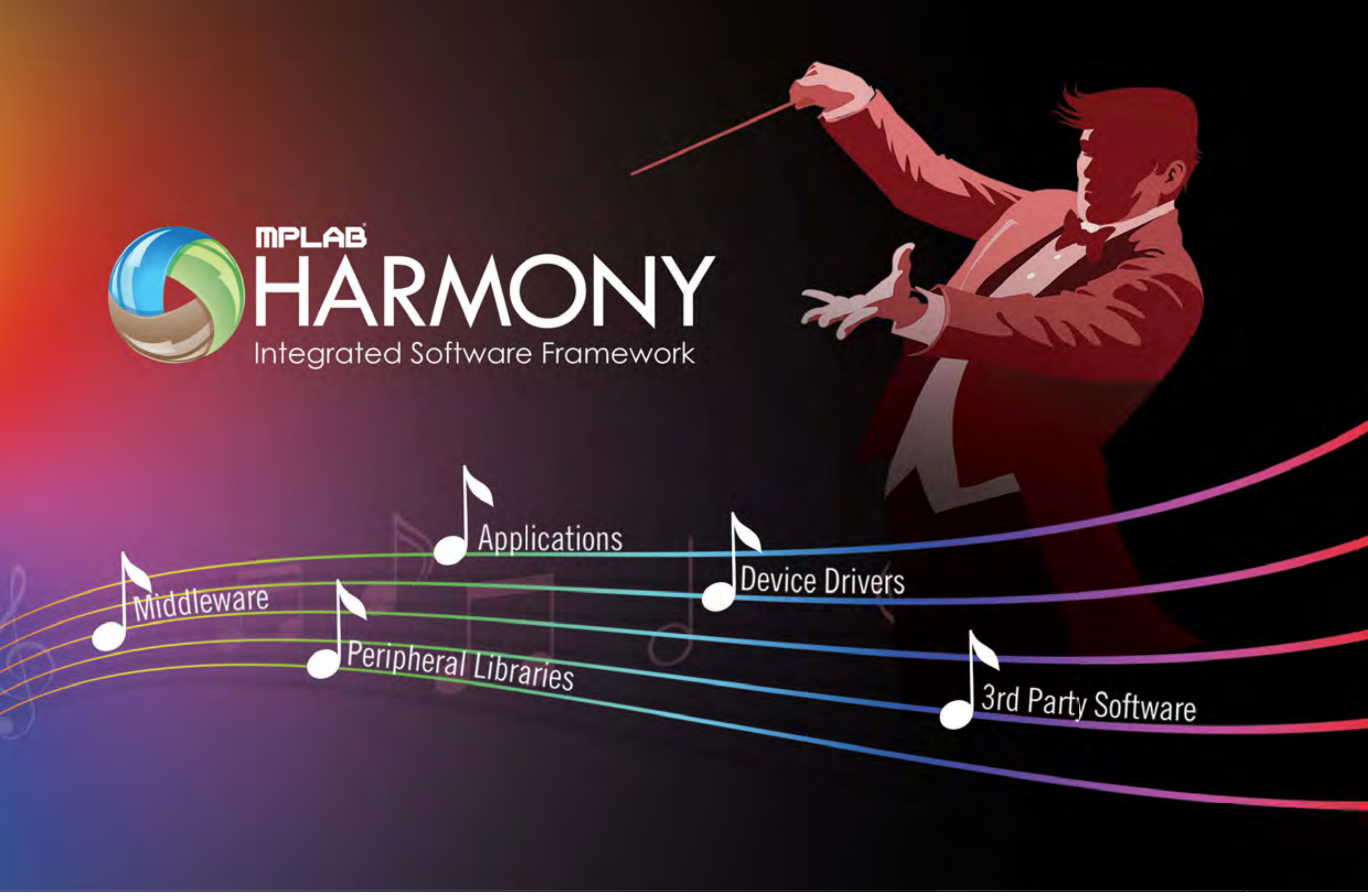
If any function throws an error, the program flow switches immediately to the event handler, where the error is handled as if it had just been popped out of the queue. There is no need to finish processing the current event or to wait for any pending events.

The event processing architecture provides a very robust and reliable framework for the *Touch-Screen Recorder* firmware and allows the addition of new functions with minimal chance of 'breaking' anything.

Next month, we will give the assembly details, provide some performance graphs and show how to use and set up the *Touch-Screen Recorder.*

Be the Conductor!

Faster PIC32 Development with Fewer Resources



Code Interoperability

- Modular achitecture allows drivers and libraries to work together with minimal effort

Faster Time to Market

- Integrated single platform enables shorter development time

Improved Compatibility

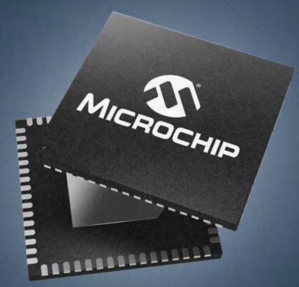
- Scalable across PIC32 Microchip parts to custom fit new project requirements

Quicker Support

- One stop support for all of your design needs, including third party solutions

Easy Third Party Sofware

- Integrates third party solutions into the software framework seamlessly
  - RTOS
  - Middleware
  - Drivers, etc



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TechnoTalk

A miss is as good as a mile

Mark Nelson

In both 2013 and 2014 the media reported stories about near-Earth asteroids that could have wrought havoc by smashing into our planet. In fact, they missed us. Last year our electronics hobby was also at risk when a European Commission directive announced tougher controls that could have regulated home construction activities. Fortunately, it appears that once again we have been spared. Mark Nelson breathes a sigh of relief.

WE ELECTRONICS ENTHUSIASTS and radio amateurs are remarkably lucky in having low 'visibility' to the authorities and the media. This happy state of affairs means we can normally enjoy our hobby without attracting adverse comment or control. I can only presume that 'they' see us as relatively harmless, posing no great threat to society, which has to be an ideal state of affairs. We can construct mains-powered devices, we can attach telephonic devices to the public telecommunications network and even transmit radio waves on certain defined frequencies, all without the need for explicit permission or third-party inspection. We can self-certify our construction projects and do not have to attach 'Chinese Export' (CE) labels to things we make for ourselves. Hallelujah for that!

RED tape by the box load

All this is thanks to an agreeable absence of red tape, although we might well have suffered a dose of RED tape, leaving you seeing RED. Forgive these two puns but 'RED' is highly relevant here. RED is the Radio Equipment Directive, enacted by the European Parliament last year, and coming into force in June 2016. Member states of the European Community have just one year to incorporate this directive into national law, with manufacturers being given an extra year to comply with the new regulations.

It's a substantial document (45 pages no less!) and its aim is to correct the current low level of compliance and conformity with the rules. It is also wide-ranging, as Jean-Louis Evans of testing and certification company Tüv Süd explains. 'All radio equipment is subject to the RED. 'Radio equipment' is defined as an electrical or electronic product that intentionally emits or receives radio waves for the purpose of radio communication and/or radio determination, or an electrical or electronic product that includes an accessory (such as an antenna) so as to intentionally emit and/or receive radio waves for the purpose of radio communication and/or radio determination,' he states.

Get Out of Jail card for hobbyists

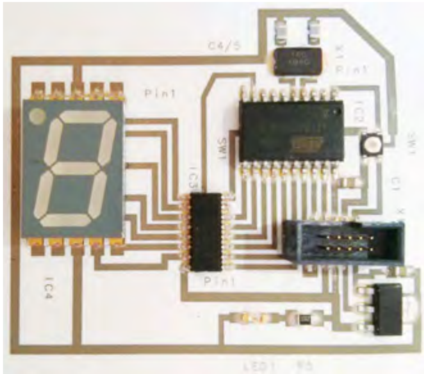
Although the RED covers all radio and TV receivers, mobile devices and wireless applications, there are some exclusions. It does not cover equipment that is used for law enforcement, public security, defence and state security. Also excluded are marine and aeronautical radio equipment, also custom-built evaluation devices designed for professionals to use at R&D facilities. Fixed-line telecom terminal equipment, such as cordless phones is now outside the scope of the RED and is instead covered by the EMC and Low Voltage Directives. One final exclusion is equipment used by radio amateurs and while this does not alter the current state of affairs, it is still welcome news. As for wireless devices constructed by hobbyists, so long as they operate on licence-exempt frequencies, these too escape the new RED rules, so let's be grateful for small mercies.

Print your own PCBs

For years people have been etching their own printed circuit boards at home, with varying degrees of success. But now there's a new device that does far more than that, creating a PCB factory all in one simple machine that fits on your desk. It probably won't fit your wallet – yet – but just look how the cost of desktop 3-D plastic printers has plummeted in a relatively short time.

In fact, the Squink (that's the name of this new product from BotFactory) looks uncannily similar to a 3-D printer, at least at first sight. Its capabilities are three-fold. First, Squink prints conductive ink on a surface to create the tracks and pads of the circuit, using standard inkjet technology. You can print Gerber RS-274X files or upload PNG, JPG or BMP files. According to BotFactory, the intuitive web interface will guide you through the process, letting you print complex circuits in a matter of minutes.

The next step is even more revolutionary. Squink avoids the need of soldering components in place by using the soldermask file generated by your CAD tool to place dots of *conductive* glue at every spot where a part needs to be connected. Last comes the pick-and-place process.



Squink circuit (image courtesy of BotFactory)

Squink is designed for assembling PCBs using SMD components and uses a vacuum sucker tool to pick components from a tray. Then it aligns them using computer vision, rotates them according to the 'Centroid and Rotation' file created in your CAD tool and places them accurately. Each component is picked from a tray, not a rail, so the setup of the machine is as simple as it gets, says BotFactory. Up to 20 slots can be configured per job, allowing for automated assembly without the complexity of large industrial machines. The total time for assembling a 4 × 4-inch board with 15 components is quoted as 30 minutes.

OK, so what's the cost? Around £2,350, which will almost certainly fall over time. The project was successfully funded on Kickstarter, and if you go to YouTube you can see a whole host of videos showing the device in action on a variety of constructional projects. The BotFactory website is at: **https://www.botfactory.co** (*no* 'UK' after '.co') and I have to say, I find it utterly fascinating! But what about multi-layer boards? Can Squink handle them? No, but Voltera can, as well as letting you use the printer to dispense solder paste to outsourced boards before picking and placing. It also includes a reflow oven. Like Squink, Voltera used crowdfunding to raise its capital cost. Its goal was $70,000 but the amount raised already is approaching $500,000. Take a look at **http://voltera.io** and **https://www.facebook.com/voltera.io** to see how quickly one innovation can inspire improved copies! Everybody will be making these machines soon…

Raspberry Pi for Dummies

Mike Tooley takes a break from Teach-In 2015 to explore one of the latest titles in the long-running and immensely popular series of 'Dummies guides'.

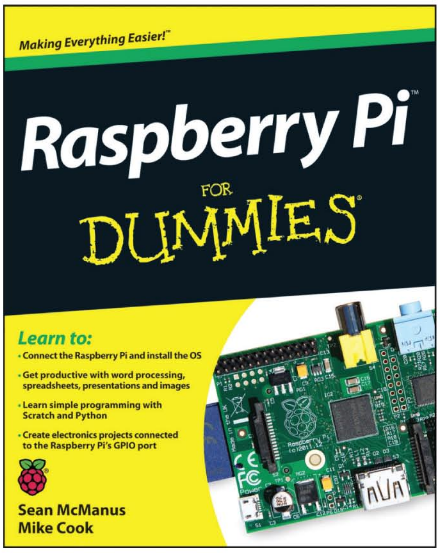
Anything but Dumb!

Established in 1969, the *Dummies* guides have become a well-established and highly respected source of information on topics varying from *Astronomy* to *Excel VBA Programming* and from *Beating Sugar Addiction* to *Gardening with Free-Range Chickens*. This new edition of the *Raspberry Pi for Dummies* guide adds to this eclectic collection by providing a comprehensive reference for anyone who is new to the Pi or who has one and wants to make it do more. In fact, there's something in this book for everyone, not just newcomers. Established users will find plenty to get their teeth into – but, a word of caution might be appropriate here, as the 'electronic content' of this book is rather basic and many *EPE* readers might find that the book does not go quite far enough.

Sections

The book comprises over 400 pages arranged in 20 chapters. A welcome addition to this latest edition is coverage of the recently introduced Model B+. The content is arranged in six parts, plus appendices and a bonus on-line chapter. Part I – 'Getting Started with the Raspberry Pi' – will provide you with a useful introduction to the Pi, its operating system, downloading software and how to configure the Pi for first-time operation. This is all valuable information, particularly for new users.

Part II covers 'Getting Started with Linux'. It provide an overview of the operating system, the desktop environment, and using the Linux shell. I suspect that for most *EPE* readers this will be extremely useful. Part III is titled 'Using the Raspberry Pi for Both Work and Play' and deals with using an office application, photo editing, and using the Pi with media. Most *EPE* readers are likely to find this of somewhat peripheral interest, but rather more usefully, 'Programming the Pi' is covered in Part IV. This deals with using Scratch, Python, Minecraft and making music. With the exception of the chapter on Python (about 30 pages) this is also likely to have a somewhat limited appeal to the average electronics enthusiast.



Electronics

Part V moves on to 'Exploring Electronics with the Raspberry Pi', and I suspect this is where the book will start to get interesting for most *EPE* readers. This part begins with a chapter on 'Understanding circuits and soldering' and, while this might be very important for newcomers, it is rather too basic for most *EPE* readers. This is followed by a chapter devoted to making a breakout board. Information is given on preparing ribbon cable, using a vice to fit an IDC connector to a ribbon cable, and terminating the ribbon cable on two rows of rather chunky screw terminal block connectors. There are better, neater and more reliable ways of doing this, such as those available in both kit and ready-assembled form from: **https://www.modmypi.com**.

The remainder of the chapter describes the construction of a 'Blastoff' game, which uses the Raspberry Pi and GPIO breakout box to sense the presence of a marble at various points on a board. The following chapter introduces a 'Copycat' game using LEDs and push-button switches mounted on stripboard. The author goes on to explain how a transistor can be used to increase the (limited) output current drive capability of the Pi's GPIO port and, although this builds into a rather nice little project, it hardly justifies the title 'Raspberry Pi in control' (which might suggest a much wider range of I/O applications, not just switches and LEDs).

Fortunately, the next chapter delves into the analogue world with the construction of a simple analogue input/output board and concludes with some interesting projects, including a curve tracer and temperature sensor. Code for use with these (and other) projects is available for download from the book's companion website.

Part VI is oddly called 'The Part of Tens' (a title that alludes to the ten software packages and ten inspiring (sic) projects that are described in this part). The information given is rather brief, but the aim of this section is to provide ideas, projects and sources of information. The book concludes with some useful appendices and a comprehensive index.

Summary

Priced at £17.99, and bearing in mind the wide-ranging content, this book represents good value and is a worthwhile addition to the bookshelf of Raspberry Pi users. Experienced electronics enthusiasts will probably find some of the 'electronic content' rather basic, but the chapters on the desktop environment, Linux and Python will at least provide some compensation. Least useful is probably the chapter on using Scratch to develop an arcade game (about 20 pages) and programming **Minecraft.Pi.html**

Book details

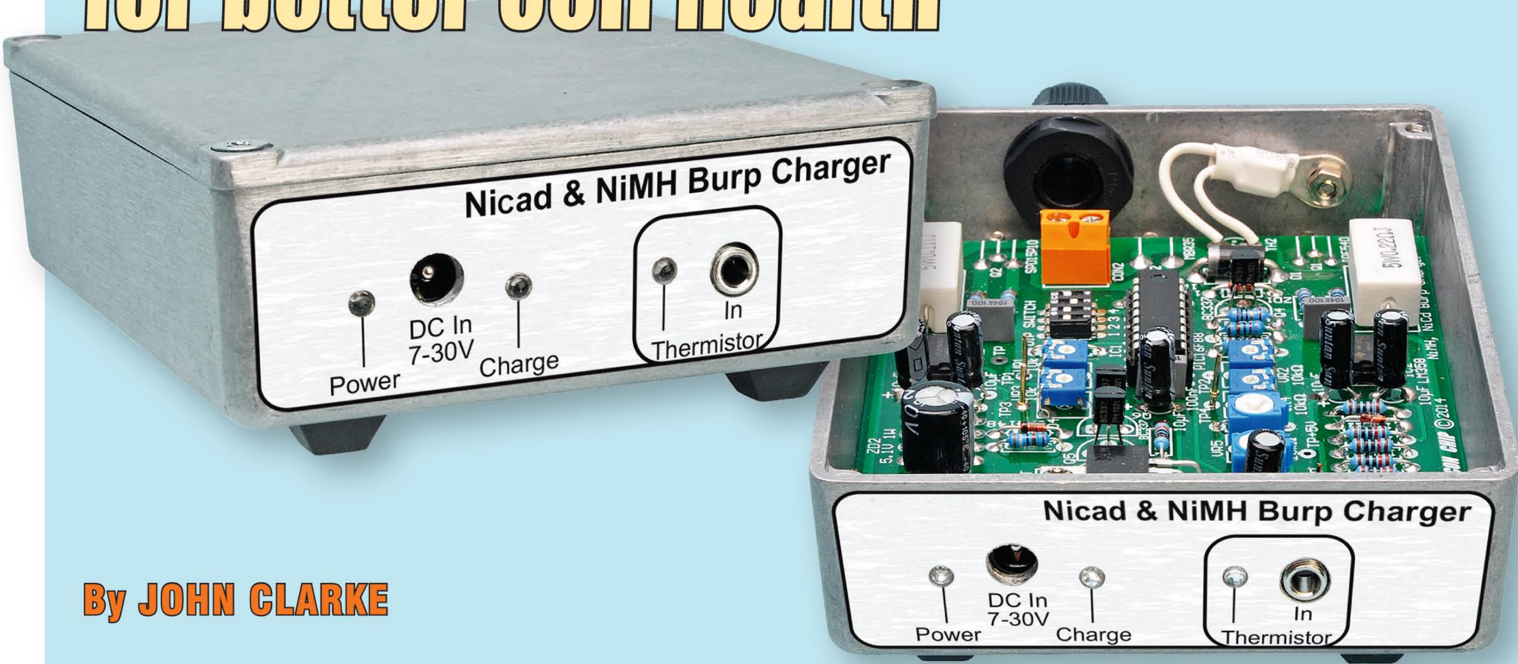
Raspberry Pi for Dummies by Sean McManus and Mike Cook

John Wiley and Sons
ISBN 978-1-118-90491-6

Raspberry Pi for Dummies is available from EPE Book Services (Reference RPiDUM01). Further details can be found at: www.epemag.wimborne.co.uk/acatalog/Raspberry_Pi.html

Burp Charge Your Batteries

for better cell health



By JOHN CLARKE

Most readers know that Nicad and NiMH batteries can be fast charged, but an even better way of doing it is to 'burp' charge them. This is a rapid alternate charge and discharge process that reduces pressure and temperature build-up in the cells, and as a result, increases the charging efficiency.

THIS VERY versatile *Nicad and NiMH Burp Charger* can charge a single cell or up to 15 series-connected cells. All the standard charge cycles, such as fast charge, top-up and trickle are available, together with the added benefits of burp charging. Built-in safeguards include temperature sensing of the cells to prevent overcharging, as well as sensing inside the charger itself for over-temperature protection.

The concept of burp charging has been known since the late 1960s. At that time though, the many benefits thought to be associated with it were largely unsubstantiated. A specialised IC (the ICS1702) was developed that incorporated burp charging (see: **www.klaus-leidinger.de/mp/RC-Elektronik/Reflexlader/ics1702.pdf**) and this became the basis for commercial burp chargers and for chargers used by NASA for Nicad cells in space applications. However, these chargers were used without any real understanding as to why burp charging was beneficial.

It wasn't until 1998 that an exhaustive investigation compared standard charging with burp charging in a research paper entitled *Investigation of the Response of NiMH Cells to Burp Charging* by Eric Darcy (see **http://corsair.flugmodellbau.de/files/elektron/NASA-II.PDF** or in condensed form at **ntrs.nasa.gov/search.jsp?R=20000086665**). This research proved that burp charging improved cell performance compared to other charging techniques.

It was found that the burp process caused the oxygen bubbles produced during charging to be re-absorbed back

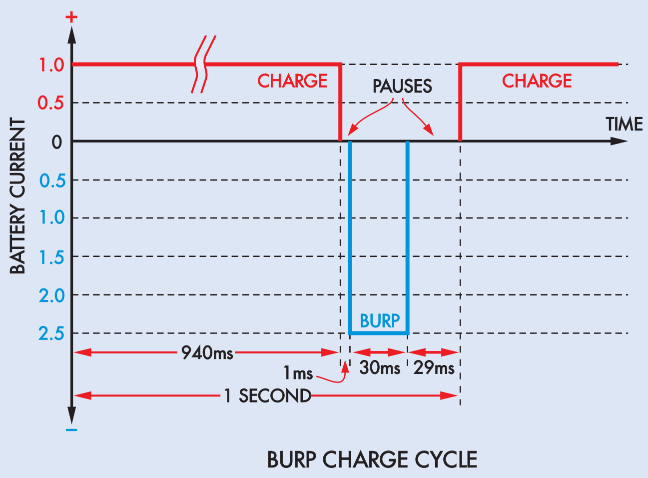


Fig.1: the charge, pause and discharge (burp) cycles for our *Nicad & NiMH Burp Charger*. It comprises a 940ms charge period followed by a 1ms pause, then a 30ms discharge period, followed by a 29ms pause, giving a total cycle of one second.

in the electrochemical process. With oxygen levels lowered, there is less pressure build up inside the cell. In addition, the lack of oxygen bubbles increases the available surface area on the cell electrodes and results in more efficient charging.

The research also found that while many commercial burp chargers, including those that use the ICS1702 IC, used a 5ms discharge (burp) period, a period of 30ms was more beneficial. That's because a longer discharge period allows more complete recombination of generated oxygen. For that reason, a 30ms burp period is used in the new design described here.

By the way, the term 'burp' charging is something of a misnomer as the oxygen is not 'burped' or released. Instead, it is recombined or consumed at the positive electrode surface.

Charge/discharge cycle

Fig.1 shows the sequence of charge, pause and discharge (burp) for the *Nicad & NiMH Burp Charger*. It comprises a 940ms charge period, followed by a 1ms pause, then a 30ms discharge period, followed by a 29ms pause, giving a total cycle of one second (1s). On this figure, a charging period is shown as having a value of '1', while a discharge (burp) period is assigned a value of '–2.5'. This simply means that the discharge current is 2.5 times the charge current.

Burp chargers are not commonly available, but standard NiMH/Nicad chargers can be obtained just about anywhere. However, the latter usually only charge two or four AA cells at a time and they charge at quite a slow rate, typically taking 4-15 hours for a full recharge.

But what if you want to charge at a much higher rate, or you want to charge more than four cells at a time, or if you want to use burp charging? Or what if you want to cater for 'C' and 'D' cells or battery packs? The answer is to build our *Nicad & NiMH Burp Charger*.

This new unit can charge from 1-15 NiMH or Nicad cells simultaneously; ie, battery packs up to 18V. In addition, the charging rate can be set from just a few milliamps up to 2.5A and it includes reliable end-of-charge detection (using temperature sensing), with extra safeguards to prevent over-charging.

Safety is important when charging NiMH and Nicad cells because they can have their life seriously shortened if the charger is left on for too long after the battery pack has reached full charge. Worse still, the cells can be destroyed or explode if over-charged.

To see why over-charging can destroy a battery pack, take a look at Fig.2. This shows the typical voltage, temperature and internal pressure rise of a cell or battery pack during charging. Once charging goes past the 100% point, the temperature and internal pressures rapidly rise, while the voltage initially rises and then falls.

Continual overcharging will damage the cells due to the elevated temperature. This accelerates chemical reactions that contribute to the cell's ageing process. In extreme cases during overcharging, excessive internal pressure can open the safety vents to release the pressure. These vents will then re-close after the pressure has been released, but by that time the cells will already have been damaged.

Full charge detection

Full charge can be determined in one of two ways. The conventional way has been to monitor the voltage across the battery pack and detect the point at which the voltage suddenly begins to rapidly rise and then fall. This form of charge end-point detection is called dV/dt (ie, change in voltage with respect to time).

In practice, the critical end-point can be difficult to detect at low currents, particularly with NiMH cells.

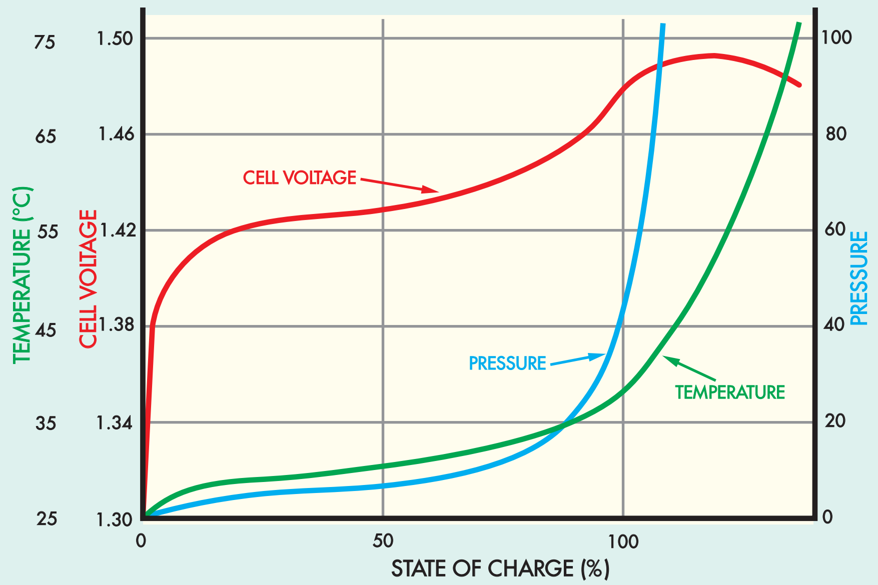


Fig. 2: typical charging curves for NiMH/Nicad batteries. Cell temperature (green) and voltage (red) rates of change are often used to detect the 'end point' (100% charge), although voltage-rate detection is not reliable in NiMH cells.

Main features

- Designed for charging NiMH and Nicad cells
- Optional top-up and trickle charging
- Optional burp charging
- Adjustable charge current
- Charging time-out
- dT/dt (temperature change rate) for end of charge detection
- Over and under cell-temperature detection
- Power, charging and temperature indication LEDs
- Adjustable charging time-out limit
- Adjustable dT/dt setting
- Adjustable top-up and trickle charge currents
- Over-temperature cut out for charger

In fact, dV/dt end-point detection with NiMH cells is neither safe nor practical. The only safe way is to monitor the temperature of the cells, but very few chargers do this.

Basically, this latter method of end-point detection monitors the temperature rise of one or two cells within the battery pack. During charging, the cells do not heat up much because most of the incoming power is converted into stored energy. However, once the cells become fully charged, the charging power is converted to heat and so the cells quickly rise in temperature.

This temperature change at the charging end point is called dT/dt, ie, change in temperature over time. The critical rate is of the order of 2°C per minute, and this is the point where normal charging should cease.

Some chargers, this one included, include a top-up charge after the end-point to ensure full charging. The top-up charge rate is less than the main charge current and is set at four times the trickle current setting.

Finally, after the top-up cycle, the cells can be trickle-charged at low current to maintain full charge. In this situation, the cells are deliberately left connected to the charger so they are fully charged when needed.

Our new burp charger monitors the cell temperature using a small thermistor. This is installed in the battery pack or cell holder, in close contact with one of the cells. The beauty of this system is that it will reliably detect the end of charge (end-point) of any type of cell, regardless of whether it was initally completely flat or only partially discharged.

Note that when charging very cold batteries, there may be a rapid rise in temperature during charging. This could cause a false dT/dt end of charge indication. To circumvent this, the dT/dt measurement for end of charge detection is only enabled when the cell temperature is at least 25°C. Should the thermistor end-point detection fail, a timer is included that will switch off charging after a preset period.

Further safeguards to protect the cells are also included. For example, charging will not start or will cease if the NTC thermistor is disconnected or if the temperature is under 0°C or over 50°C. In addition, if the charger itself becomes too hot, charging will pause and the temperature is measured after two minutes to check if it has cooled sufficiently to restart.

Select the features you want

In its simplest form, the charger includes only the temperature detection feature, after which charging ceases. However, you can add top-up and trickle charging if you want. In addition, the charging rate can be set for both the main charge current and the trickle charge, along with the time-out period and dT/dt values.

In practice, the main charging rate can be set from about 40mA up to 2.5A, while trickle-charging can be set from 10-500mA. The time-out can be set from between 0-25 hours, while dT/dt can be selected from between 0.5°C per minute to 5°C per minute.

Further details concerning these adjustments are included in the setting-up section of this article.

Three front-panel LEDs are used to indicate the status of the charger. First, the Power LED is lit whenever power is applied to the charger, while the Thermistor LED lights if the thermistor is disconnected or if there is an over or under-temperature detection. For over-temperature (>50°C), the Thermistor LED will flash once a second (1Hz) while for under temperature (<0°C), the LED will flash once every two seconds (0.5Hz).

Over-heating of the charger itself causes the Thermistor LED to flash once every four seconds.

Finally, the Charging LED is continuously lit during the main charging cycle and switches off when charging is complete. If top-up or trickle charging are selected, the charging LED will flash at 1Hz during top-up and at 0.5Hz during trickle charge (ie, at 1s and 2s intervals respectively).

Note that if the Thermistor LED is lit or flashing, the charging LED will be off, indicating that charging has either paused or ceased.

Circuit details

Now take a look at Fig.3 for the circuit details. It's based on IC1, a PIC16F88-I/P microcontroller, plus MOSFETs Q1 and Q2. Q1 is used for charging, while Q2 is used for the burp discharging.

In addition, two NTC thermistors, TH1 and TH2, are used. TH1 monitors the temperature of the cell or battery pack being charged. It's connected via a 3.5mm jack plug and socket (CON3) and together with 20kΩ trimpot VR5, forms a voltage divider across the 5V supply. VR5 is adjusted so that the voltage across the thermistor is 2.5V at 25°C (note: NTC stands for 'negative temperature coefficient' and means that the resistance of the thermistor progressively reduces as the temperature rises).

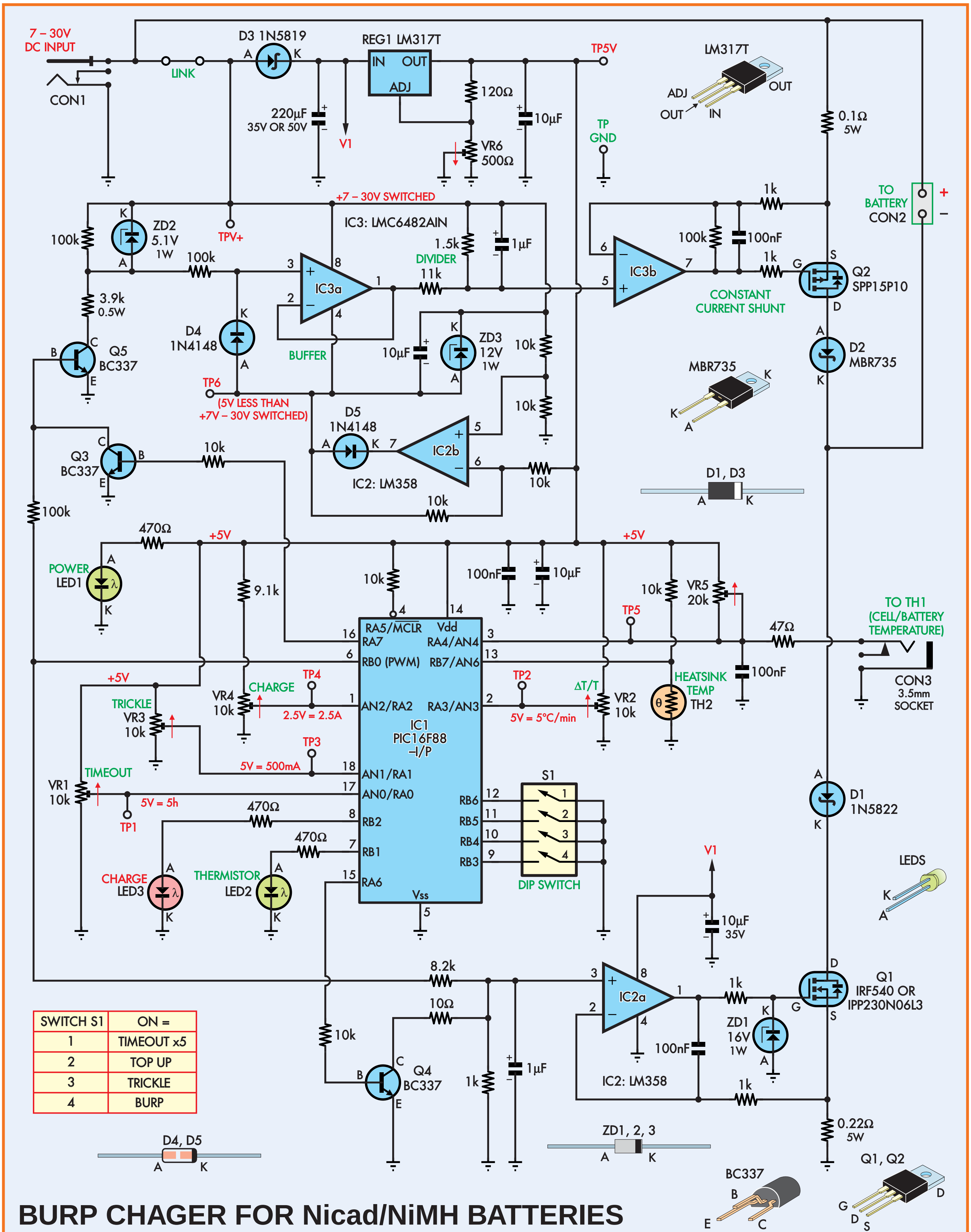
The voltage across TH1 is monitored at the AN4 input (pin 3) of IC1 via a 47Ω resistor and 100nF filter capacitor. These are included to remove RF (radio frequency) signals and noise that could be present due to the thermistor being connected remotely from the circuit. The voltage at the AN4 input is then converted into a digital value and monitored for dT/dt changes. It is also compared by IC1 against stored over and under-temperature values.

TH2 is connected to the AN6 input of IC1 and monitors the charger's heatsink temperature. This allows IC1 to shut the charger down if the heatsink temperature exceeds a preset value.

Trimpots VR1, VR2 and VR3 are used to set the time-out, dT/dt and trickle charge values. These trimpots connect to AN0, AN3 and AN1 of IC1 respectively and are be set to apply between 0V and 5V to these inputs.

Trimpot VR4 sets the charging current. This trimpot connects to the +5V supply via a 9.1kΩ resistor and this restricts the adjustment range to a nominal 2.5V maximum at IC1's AN2 input (pin 1), corresponding to a 2.5A maximum charge rate.

The voltage inputs are all converted to digital values within IC1 so that the



SWITCH S1	ON =
1	TIMEOUT x5
2	TOP UP
3	TRICKLE
4	BURP

BURP CHAGER FOR Nicad/NiMH BATTERIES

Fig.3: the circuit is based on IC1. This accepts inputs from TH1 and TH2, trimpots VR1-VR5 and DIP switch S1, sets the charge rates and the time-out, and controls the charging current through Q1 via its PWM output (RB0). IC1's PWM output also drives Q5 and IC3a, which then drive a current shunt based on IC3b and Q2 to provide the discharge circuit.

Parts List

1 PCB, available from the *EPE PCB Service,* code 14103141, 105 × 87mm
1 119 × 94 × 34mm diecast case
1 2.5mm DC socket (CON1)
1 3.5mm stereo PCB mount jack socket (CON3)
1 3.5mm mono line jack plug
1 2-way screw terminal, 5.08mm spacing (CON2)
1 4-way DIP switch (S1)
2 DIL 8-pin sockets (optional)
1 DIL18 IC socket
2 10kΩ @ 25°C NTC thermistors (TH1,TH2)
1 crimp eyelet, 5.3mm ID (to mount TH2)
3 TO-220 silicone insulating washers
3 TO-220 insulating bushes
1 cable gland for 3-6.5mm cable
4 rubber feet
4 M3 × 6.3mm tapped spacers
8 M3 × 5mm screws
5 M3 × 10mm screws
5 M3 nuts
1 M3 star washer
1 200mm length of single-core screened cable
7 PC stakes
Hook-up wire, heatshrink, etc

Semiconductors
1 PIC16F88-I/P microcontroller programmed with 1410314A.hex (IC1)
1 LM358 dual op amp (IC2)
1 LMC6482AIN CMOS dual op amp (IC3)
1 LM317T adjustable regulator (REG1)
1 IRF540 or IPP230N06L3 N-channel MOSFET (Q1)
1 SPP15P10PLH P-channel logic level MOSFET (Q2)
3 BC337 NPN transistors (Q3-Q5)
1 1N5822 3A Schottky diode (D1)
1 MBR735 7A Schottky diode (D2)
1 1N5819 1A Schottky diode (D3)
2 1N4148 diodes (D4,D5)
1 16V Zener diode 1W (ZD1)
1 5.1V Zener diode 1W (ZD2)
1 12V Zener diode 1W (ZD3)
2 3mm green LEDs (LED1, LED2)
1 3mm red LED (LED3)

Capacitors
1 220µF 35V or 50V PC electrolytic
4 10µF 35V or 50V PC electrolytic
2 1µF 16V PC electrolytic
4 100nF 63V or 100V MKT polyester

Trimpots
4 10kΩ horizontal trimpots (VR1-VR4)
1 20kΩ horizontal trimpots (VR5)
1 500Ω horizontal trimpot (VR6)

Resistors (0.25W, 1%)

4 100kΩ	5 1kΩ
1 11kΩ	3 470Ω
8 10kΩ	1 120Ω
1 9.1kΩ	1 47Ω
1 8.2kΩ	1 10Ω
1 3.9kΩ 0.5W	1 0.22Ω 5W
1 1.5kΩ	1 0.1Ω 5W

settings can be processed in software. Test points TP1-TP5 are provided for setting the trimpots when using a multimeter. There is also a TP GND terminal which is useful when checking these voltages.

The voltages measured at each test point directly relate to the setting's value. For example, setting VR1 to give 4V at TP1 will set the time-out to four hours. This time-out value can be multiplied by a factor of five by setting the No.1 switch in DIP switch S1 to the ON position. This ties pin 12 (RB6) of IC1 to ground.

Conversely, with this switch open, pin 12 is pulled to +5V via an internal pull-up resistor within IC1 and the time-out is set to ×1. Switches 2, 3 and 4 in DIP switch S1 work in a similar manner. The No.2 switch enables the top-up, the No.3 switch enables the trickle-charge mode and the No.4 switch enables the burp charge.

Outputs RB1 and RB2 of IC1 drive the Thermistor and Charge indicator LEDs (LED2 and LED3) respectively via 470Ω resistors. These indicate the charger's status, as described previously.

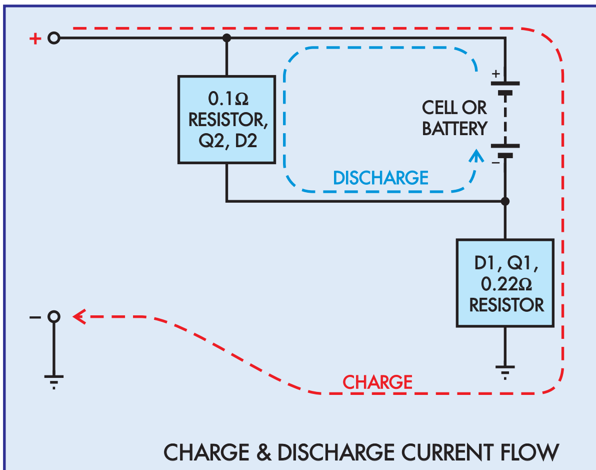


Fig.4: the basic charge and discharge current paths for the unit. During charging, current flows from the power supply, through the cell or battery and then via diode D1, MOSFET Q1, and a 0.22Ω resistor to ground. Conversely, during discharge, current flows from the cell or battery through MOSFET Q2, diode D2 and a 0.1Ω resistor.

Charge and discharge

Two separate circuits are used for the charge and discharge functions. To understand how this works, refer to Fig.4, which shows the basic charge and discharge current paths.

During charging, current flows from the power supply through the cell or battery and then via diode D1, MOSFET Q1 and a 0.22Ω resistor to ground. Conversely, during discharge, current flows from the cell or battery through diode D2, MOSFET Q2 and a 0.1Ω resistor. Note, however, that this is a simplified diagram and the currents through Q1 and Q2 are controlled so that charge and discharge rates are correct for the cell or battery that's connected to the charger.

Refer back now to Fig.3 for the full details. A constant current source comprising op amp IC2a and MOSFET Q1 charges the battery via CON2. IC1's RB0 output provides a 5V 3.9kHz PWM (pulse-width modulated) signal which is fed to a divider and filter

network comprising 8.2kΩ and 1kΩ resistors and a 1μF capacitor. This filter network smooths the pulse output to give a DC voltage.

This smoothed DC voltage sets the current provided by Q1 to the battery and the 5V PWM signal has its duty cycle adjusted over a wide range, from trickle to full charge. The 5V level is effectively reduced to 543mV via an 8.2kΩ and 1kΩ voltage divider. As a result, the maximum voltage that can be applied to pin 3 of IC2a is 543mV when the PWM duty cycle is 100% (ie, full charge). For a 50% duty cycle, the average voltage from RB0 is 2.5V, or 271.5mV after passing through the divider.

This filtered voltage is applied to pin 3 of IC2a and this sets the charge current. When pin 3 is at 543mV, IC2a's pin 1 output adjusts the gate drive to MOSFET Q1 so that the voltage across the 0.22Ω source resistor (as monitored at pin 2 of IC2a) is also 543mV. The charge current is therefore 2.47A (ie, 543mV ÷ 0.22Ω).

Diode D1 is included to prevent current flow via Q1's intrinsic reverse diode if power is connected with reverse polarity. D1 is a 3A Schottky type, specified because it has less than half the forward voltage of a normal power diode. Typically, it has about 380mV across it (at 2.5A) compared with a standard diode which would have 0.84V across it at 2.5A.

That also means less power loss in the diode; 0.95W for the Schottky diode compared to 2.1W in a standard diode.

IC1's RA6 output drives transistor Q4. This transistor is used to pull the voltage at pin 3 of IC2a to a very low level, so that the charge current is effectively reduced to near zero. This shut down is required during pause (when the PWM is also dropped to zero) and also during discharge when the PWM is still present to provide the discharge current setting.

Specifications

- Maximum input voltage: 30V.
- Maximum charge current: 2.5A.
- Charge current adjustment: from 0-2.5A, corresponding to 0-2.5V at TP4 using VR4 (in approximately 40mA steps).
- Time-out adjustment: from 0-5 hours, corresponding to 0-5V at TP1 using VR1. 0-25 hours with x5 selected (when DIP switch 1 closed).
- dT/dt adjustment: from 0.5-5°C rise/minute, corresponding to 0.5-5V at TP2 as set by VR2.
- dT/dt measurement interval: once every minute when cells reach 25°C or more.
- Top-up and trickle charge: top-up available when DIP switch 2 is closed; trickle enabled when DIP switch 3 is closed.
- Trickle charge: adjustable using VR3 from 0-500mA, corresponding to 0-5V at TP3. Adjustable in approximately 5mA steps.
- Top-up charge: 4 x trickle setting for one hour.
- Burp discharge: enabled when DIP switch 4 is closed. Discharge current is 2.5 times the charge current. Time-out is increased by 13% to compensate for reduced charge period and added discharge period.
- Cell over-temperature cut-out: 50°C.
- Cell under-temperature cut-out: 0°C.
- Charger over-temperature cut-out: 40°C.
- Charging cycle with burp selected: charge period 940ms, pause 1ms, burp discharge 30ms and pause 29ms (all over a 1s period).

Burp discharge

Another constant current circuit is employed for the burp discharge function. This comprises op amp IC3b and MOSFET Q2, with a 0.1Ω source resistor used for current monitoring. This circuit is connected to the positive supply (instead of the 0V supply as for the charge circuit) and so Q2 is a P-channel MOSFET. In addition, the PWM signal for IC1 is inverted and referenced to the positive supply.

The same PWM signal from RB0 (pin 6 of IC1) is also used to control IC3b and Q2. However, because we now have a P-channel MOSFET, the signal is inverted and level-shifted by transistor Q5. When the PWM signal is at 5V, Q5 switches on and its collector goes low, pulling one side of the 3.9kΩ resistor low. This 3.9kΩ resistor limits the current flow through 5.1V Zener diode ZD2. This Zener diode clamps the inverted voltage to within 5V of the switched supply rail.

As a result, the 5V PWM signal is now inverted and referenced below the positive supply which can be as high as 30V, depending on the number of cells being charged.

IC3a is powered from a 5V supply; ie, between the 30V positive rail at its pin 8 and a rail 5V below this at pin 4. A 100kΩ resistor couples Q5's output to pin 3 of IC3a and this resistor limits the current into clamp diode D4. D4 prevents the voltage applied to pin 3 going much below the pin 4 rail, thereby preventing damage to this op amp.

IC3a essentially buffers the PWM signal before feeding it to op amp IC3b via an 11kΩ/1.5kΩ divider. A 1μF capacitor filters the divider's output. This divider is designed to automatically provide a discharge current that's 2.5 times greater than the charge current.

The 5V inverted PWM signal that's now referenced to the positive supply becomes a 600mV signal (again referenced to the positive supply) after the divider. When the PWM level is at maximum (ie, the charge current is 2.47A), 600mV appears across MOSFET Q2's 0.1Ω source resistor. This results in a 6A discharge current, ie, close to 2.5 times the charge current.

Power supply

Power for the circuit comes from a 7-30V DC supply (plugpack or laptop supply) via Schottky diode D3. D3 provides reverse polarity protection for the following 220μF capacitor and 3-terminal regulator REG1, an LM317T set to provide 5V to IC1 and the trimpots. This was chosen in preference to a fixed 5V regulator because it can be adjusted to supply a more precise 5V, using trimpot VR6. An exact 5V rail makes the settings of VR1-VR5 more accurate.

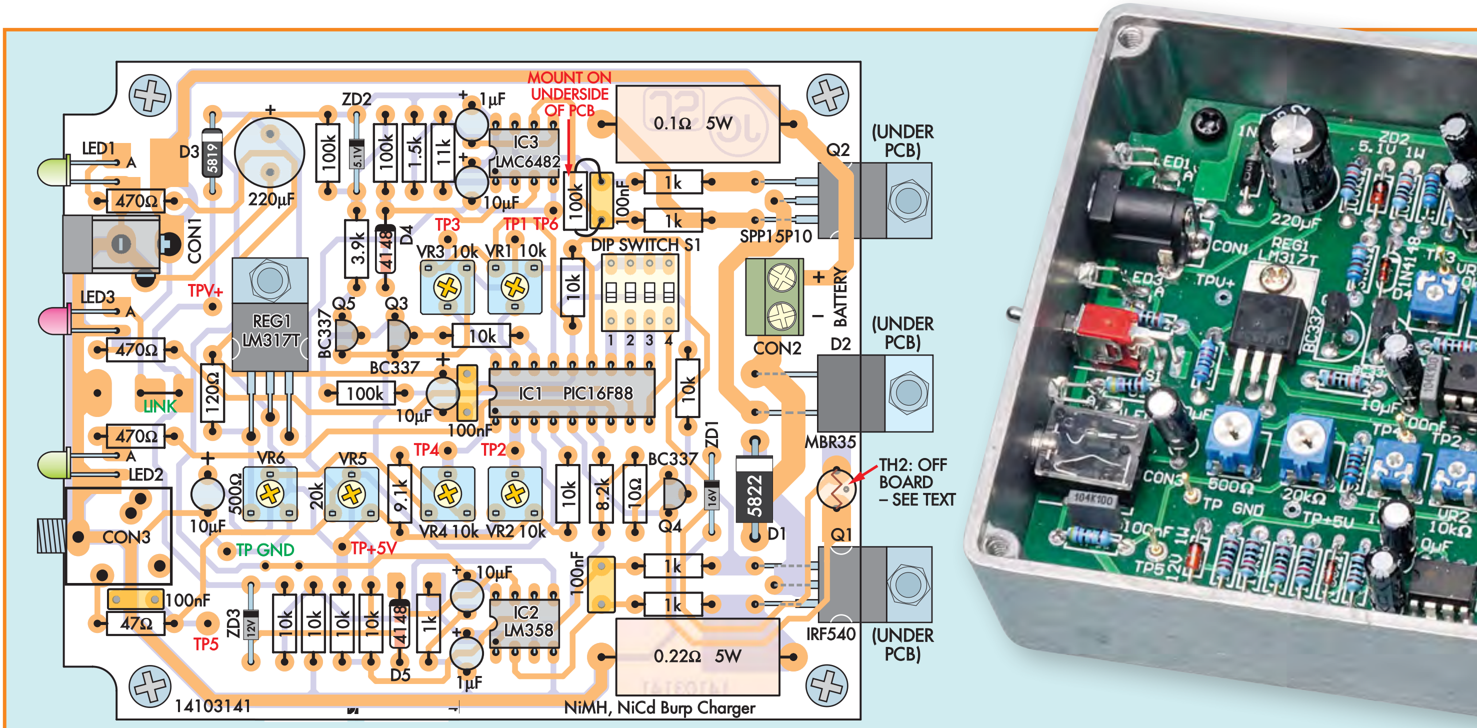


Fig.5: install the parts on the PCB as shown here. The text describes the mounting details for Q1, Q2 and D2 (see also Fig.6), thermistor TH2 and the three LEDs. Note that the 100kΩ resistor to the right of IC3 is mounted under the board, in parallel with the 100nF capacitor. In addition, the power switch shown in the photo has been replaced by a link – see panel.

The 5V supply for op amps IC3a and IC3b is provided by IC2b. This is connected to invert the 5V from REG1 and level-shift it so that it is 5V below the positive supply rail. 12V Zener diode ZD3 prevents IC2b's output from going more than 12V below the positive supply rail at power up. This protects IC3 from damage as its maximum supply rating is 16V.

D5 prevents IC2b's output from conducting current through the 12V Zener diode in the forward direction if the power supply is reversed in polarity. This also protects IC3 from damage.

Supply voltage requirements

In order to fully charge a battery, we need up to 1.8V per cell from the plugpack even though the nominal terminal voltage shown on the battery pack is 1.2V per cell. So, to charge a 6V battery which has five cells, we need a DC input voltage of 5 × 1.8V = 9V.

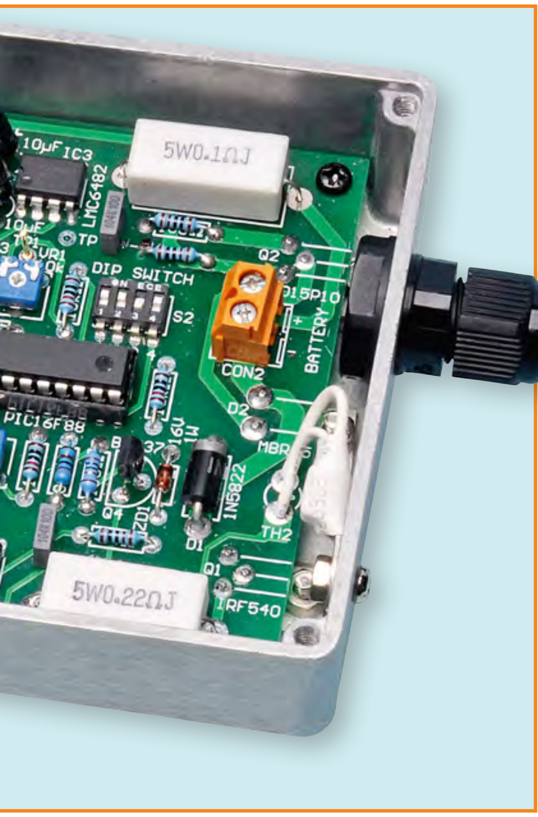
Similarly, an 18V battery has 15 cells and so this requires a 15 × 1.8V = 27V supply to fully charge it.

Charging only one, two or three cells nominally requires up to 5.4V. In practice though, more than 7V is required at the input to ensure that the LM317T regulator (REG1) operates correctly, ie, remains in regulation.

For operation in a car, the input voltage will be around 12V with the engine stopped and up to 14.4V with the engine running. A 12V supply

Table 1: Resistor Colour Codes

❑	No.	Value	4-Band Code (1%)	5-Band Code (1%)
❑	4	100kΩ	brown black yellow brown	brown black black orange brown
❑	1	11kΩ	brown brown orange brown	brown brown black red brown
❑	8	10kΩ	brown black orange brown	brown black black red brown
❑	1	9.1kΩ	white brown red brown	white brown black brown brown
❑	1	8.2kΩ	grey red red brown	grey red black brown brown
❑	1	3.9kΩ	orange white red brown	orange white black brown brown
❑	1	1.5kΩ	brown green red brown	brown green black brown brown
❑	5	1kΩ	brown black red brown	brown black black brown brown
❑	3	470Ω	yellow violet brown brown	yellow violet black black brown
❑	1	120Ω	brown red brown brown	brown red black black brown
❑	1	47Ω	yellow violet black brown	yellow violet black gold brown
❑	1	10Ω	brown black black brown	brown black black gold brown
❑	1	0.22Ω	red red silver brown	black red red silver brown
❑	1	0.1Ω	brown black silver brown	black brown black silver brown



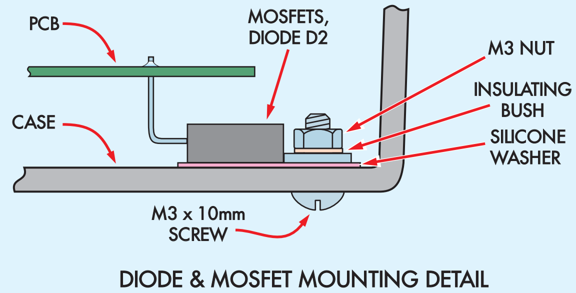


Fig.6: diode D2 and MOSFETs Q1 and Q2 are mounted on the base of the case and are insulated from it using insulating bushes and silicone washers. Make sure that the metal tab ends of the devices cannot short against the side of the case.

can charge up to six cells (ie, a 7.2V battery), while a 14.4V supply (with engine running) can charge up eight cells (ie, a 9.6V battery).

Note also that using a supply voltage that is significantly higher than required to charge the cells will cause the charger to heat up more than necessary. For example, at 2.5A and with a supply that's 10V higher than the battery voltage, around 25W will be dissipated in the charger. In that case, the charger will certainly become hot and will shut down when its heatsink (ie, the case) reaches 40°C.

Basically, this means that the charge current may have to be reduced if the supply voltage is high compared to the battery voltage. The maximum charging current is also limited by the mAh capacity of the cell or battery (see Table 2) and the rating of the DC plugpack or power supply. So in order to charge at 2.5A, the power supply or plugpack must be able to deliver this current.

Construction

The assembly is straightforward since all the parts are mounted on a PCB available from the *EPE PCB Service*, coded 14103141 and measuring 105 × 87mm. This is housed in a metal diecast case measuring 119 × 94 × 34mm.

Fig.5 shows the assembly details. Begin construction by checking the PCB for any defects such as shorted tracks, breaks in the copper and incorrect hole sizes. Also, check that the corners at the lefthand end of the PCB have been shaped to clear the internal corner posts. It's rare to find any problems but it's always a good idea to check before installing any of the parts.

Next, place the PCB inside the case and mark out the corner mounting holes in the base, noting that the PCB must sit as far to the left as it will go. This is necessary so that the 3.5mm socket later protrudes through the case side. Drill these mounting holes out to 3mm and deburr them using an oversize drill.

Now for the PCB parts. Install the small resistors first, taking care to fit the correct value in each location. Table 1 shows the resistor colour codes, but it's always a good idea to use a digital multimeter to check each one before installing it (some colours can be difficult to read).

The 0.1Ω and 0.22Ω 5W resistors can go in next. These should be mounted about 1mm above the PCB to allow air to circulate beneath them for cooling. That's easily done by pushing them down onto a 1mm-thick cardboard spacer before soldering their leads (don't forget to remove the spacer afterwards).

Next, install the diodes (but not D2), then fit IC sockets for IC1, IC2 and IC3. Be sure to orient each socket correctly, ie, with its notched end to the left. Once these are in, install the correct op amp in each position, but leave the PIC16F88 micro out for the time being.

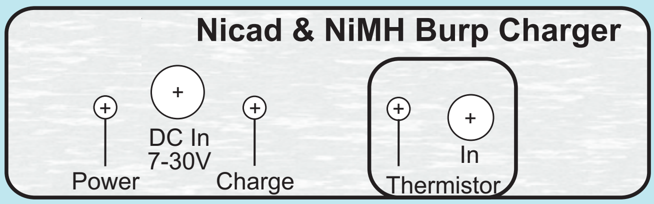
Follow with DIP switch S1, making sure that its No.1 switch goes to the left. The Zener diodes can then be installed. ZD1 is a 16V 1W type and may be marked as a 1N4745; ZD2 is 5.1V 1W and may be marked as a 1N4733; and ZD3 is 12V 1W and may be marked as a 1N4742. Again, the orientation of these parts is important.

The capacitors can now be fitted, making sure that the electrolytics go in with the correct polarity. That done, install PC stakes for TP GND, TP +5V and test points TP1-TP5.

The three LEDs are next on the list, starting with LED1 (green). First, orient it as shown on Fig.5, then bend its leads down at right angles 6mm from its body. That done, solder the LED in place with its horizontal lead sections exactly 5mm above the PCB (hint: use a 5mm-thick spacer to set the height). The remaining two LEDs can then be fitted in exactly the same manner.

Trimpots VR1-VR6 are next on the list. Note that the 10kΩ trimpots may be marked 103, the 20kΩ trimpots marked 203 and the 500Ω trimpot marked 501 (ie, instead of the actual ohm values).

Regulator REG1 is next and is mounted with its leads bent down at right angles so that its metal tab sits flat against the PCB. Secure this tab to



NOTE: POSITION LABEL SO THAT LED INDICATORS ARE 16.5mm DOWN FROM TOP EDGE OF BOX BASE

Fig.7: this full-size artwork can be used as a drilling template for the front side panel of the case.

Table 2: Typical settings for a range of cell capacities				
Battery or cell capacity	Slow charge (15h) (VR1 @ 3V, DIP Switch No.1 ON) (Do not select top up)	Standard charge (5h) (VR1 @ 5V, DIP Switch No.1 off) (Top up not recommended)	Fast charge (1.5h at or below 2.5A) (VR1 @ 1.5V, DIP Switch No.1 off)	Trickle current (DIP Switch No.3 on) (Top up with DIP Switch No.2 ON will be 4 x trickle setting)
200mAh	20mA (VR4 @ 20mV)	60mA (VR4 @ 60mV)	200mA (VR4 @ 200mV)	10mA (VR3 @ 100mV)
400mAh	40mA (VR4 @ 40mV)	120mA (VR4 @ 120mV)	400mA (VR4 @ 400mV)	20mA (VR3 @ 200mV)
700mAh	70mA (VR4 @ 70mV)	210mA (VR4 @ 210mV)	700mA (VR4 @ 700mV)	35mA (VR3 @ 350mV)
900mAh	90mA (VR4 @ 90mV)	270mA (VR4 @ 270mV)	900mA (VR4 @ 900mV)	45mA (VR3 @ 450mV)
1000mAh	100mA (VR4 @ 100mV)	300mA (VR4 @ 300mV)	1.0A (VR4 @ 1.0V)	50mA (VR3 @ 500mV)
1500mAh	150mA (VR4 @ 150mV)	450mA (VR4 @ 450mV)	1.5A (VR4 @ 1.5V)	75mA (VR3 @ 750mV)
2000mAh	200mA (VR4 @ 200mV)	600mA (VR4 @ 600mV)	2.0A (VR4 @ 2.0V)	100mA (VR3 @ 1.0V)
2400mAh	240mA (VR4 @ 240mV)	720mA (VR4 @ 720mV)	2.4A (VR4 @ 2.4V)	120mA (VR3 @ 1.2V)
2500mAh	250mA (VR4 @ 250mV)	750mA (VR4 @ 750mV)	2.5A (VR4 @ 2.5V)	125mA (VR3 @ 1.25V)
2700mAh	270mA (VR4 @ 270mV)	810mA (VR4 @ 810mV)	2.5A (1.6h) (VR4 @ 2.5V, VR1 @ 1.6V)	135mA (VR3 @ 1.35V)
3000mAh	300mA (VR4 @ 300mV)	900mA (VR4 @ 900mV)	2.5A (1.8h) (VR4 @ 2.5V, VR1 @ 1.8V)	150mA (VR3 @ 1.50V)
3300mAh	330mA (VR4 @ 330mV)	990mA (VR4 @ 990mV)	2.5A (2h) (VR4 @ 2.5V, VR1 @ 2.0V)	165mA (VR3 @ 1.65V)
4000mAh	400mA (VR4 @ 400mV)	1.2A (VR4 @ 1.2V)	2.5A (2.4h) (VR4 @ 2.5V, VR1 @ 2.4V)	200mA (VR3 @ 2.0mV)
4500mAh	450mA (VR4 @ 450mV)	1.35A (VR4 @ 1.35V)	2.5A (2.7h) (VR4 @ 2.5V, VR1 @ 2.7V)	225mA (VR3 @ 2.25V)
5000mAh	500mA (VR4 @ 500mV)	1.5A (VR4 @ 1.5V)	2.5A (3h) (VR4 @ 2.5V, VR1 @ 3.0V)	250mA (VR3 @ 2.5V)
9000mAh	900mA (VR4 @ 900mV)	2.5A (5.4h) (VR4 @ 2.5V, VR1 @ 1.08V, DIP Switch No.1 ON)	2.5A (5.4h) (VR4 @ 2.5V, VR1 @ 1.08V, DIP Switch No.1 ON)	450mA (VR3 @ 4.5V)

the PCB using an M3 × 10mm screw, nut and shakeproof washer before soldering the leads.

That done, install the DC socket (CON1), the 2-way screw terminal block (CON2) and the 3.5mm jack socket (CON3). Be sure to push these parts all the way down so that they sit flush against the PCB before soldering their leads.

That completes the PCB assembly, except for Q1, Q2 and D2. As shown on Fig.5, these three devices are all mounted under the PCB, with their leads bent up at 90° so that they pass through their respective mounting holes. This allows their tabs to be later bolted to the bottom of the metal case for heatsinking (see Fig.6).

In each case, it's simply a matter of first bending the two outside leads up by 90°, exactly 7mm from the device body. The middle leads of Q1 and Q2 can then be bent up 5mm from the body, after which you can loosely fit all three devices to the PCB – but don't solder their leads yet. **Take care not to get the two MOSFETs mixed up – Q1 is an IRF540 while Q2 is an SPD15P10.**

Case preparation

It's necessary to drill some extra holes in the case, before installing the PCB. The mounting holes for the PCB assembly were drilled in a previous step (ie, before the parts were installed) and the next step now is to use the front-panel artwork (Fig.7) as a drilling template for the front-panel holes.

You can either copy the artwork shown in Fig.7 or you can download the artwork in PDF format from the *EPE* website and print it out. In either case, it should be cut out and attached to the case using adhesive tape, after which the various holes can be drilled.

Be sure to position the label so that the centre of the LED indicators are about 16.5mm down from the top edge of the base.

Use a small pilot drill to start the holes, then remove the template and carefully enlarge each one to size using a large drill and/or a tapered reamer. There are five holes in all – three for the LEDs and one each for the DC socket and 3.5mm jack socket.

Once all the holes have been drilled, print out a final front-panel label, laminate it and attach it to the case using double-sided tape or silicone adhesive. The various holes can then be cut out with a sharp hobby knife.

Final assembly

Begin the final assembly by securing four M3 × 6.3mm tapped nylon spacers to the base of the case using M3 × 5mm screws. The PCB assembly (together with the loosely-fitted Q1, Q2 and D2

parts) can then be slipped into the case and secured to the spacers using another four M3 × 5mm screws.

The next step is to drill the mounting holes for Q1, Q2 and D2. **These devices must be positioned so that the ends of their tabs clear the side of the case by 1-2mm.** If a tab does touch the side of the case, you will have to remove the offending device and re-bend its leads so that it is clear.

Once everything is correct, remove the PCB assembly and drill the device mounting holes to 3mm, then deburr them using a larger drill. It's vital that the area around each of these holes inside the case is perfectly smooth and free of metal swarf, so that the insulating washers used when mounting the devices will not be punctured.

A hole also needs to be drilled and reamed in the adjacent side of the box (ie, at the Q1/Q2 end) to accept a cable gland (position this directly opposite CON2), while a 3mm hole must also be drilled to mount thermistor TH2. Be sure to position the hole for the cable gland down far enough so that the gland doesn't later interfere with the lid of the case.

Mounting TH2

Thermistor TH2 is attached to a 5.3mm crimp eyelet, which is then fastened to the inside of the case using an M3 × 10mm machine screw, nut and lockwasher (ie, to detect heatsink temperature).

First, remove the plastic insulating piece from the eyelet, then prise open the crimp section using pliers. That done, shape the crimp lugs so that they lightly clamp the thermistor in place but without the leads making contact to the crimp eyelet.

Finally, glue the thermistor in place using epoxy resin and heatshrink it. Then refit the PCB assembly in the case and attach the thermistor assembly to the case wall using an M3 × 10mm screw, nut and lockwasher. The thermistor's leads are then connected to its pads on the top of the PCB – see Fig.5 and photo.

Bolting down Q1, Q2 and D2

MOSFETs Q1 and Q2 and diode D2 can now be fastened to the bottom of the case. As shown in Fig.6, these devices must each be insulated from the case using a silicone washer and insulating bush. An M3 × 10mm screw and nut is used to secure each device in place, after which its leads are soldered to their pads on the top of the PCB.

Once all these devices are in, use a multimeter to check that the metal tabs of these devices are indeed isolated from the metal case. If you get a low resistance reading between a device tab and the case, dismantle the assembly and check that its insulating washer hasn't been punctured (eg, by metal swarf).

Check also that the device's tab is clear of the side of the case.

Battery-pack thermistor

As shown in Fig.8, the battery-pack thermistor (TH1) is connected to

Determining the charger settings

Before adjusting the time-out, trickle charge and time-out settings, you need to know the Ah rating (or mAh rating) of the cells or the battery. This will normally be printed on the side.

You also need to know the nominal battery voltage (or the number of cells connected in series to calculate this) and the voltage/current ratings of the plugpack.

Note that when using slow charging rates (eg, charging over 15 hours), the top-up current would exceed the charge rate. In this case, do not enable top-up. Similarly, at faster charging rates (eg, charging over five hours), the top-up current may be similar to the charge rate and again top-up is not recommended.

Charge rate

This will depend on the mAh rating of the cells or battery and on the desired charge rate (slow, standard or fast) – see Table 2. The plugpack used must also be capable of supplying the required current.

Time-out

The time-out should be set to 1.5 times the Ah rating of the battery divided by the charge current. For example, a 2500mAh (2.5Ah) battery charged at 1A should be fully charged after 2.5 hours. In this case, the time-out should be set to 2.5 x 1.5 ÷ 1 = 3.75h. That's done by adjusting VR1 to give 3.75V at TP1 (see text).

Note that any changes made to the time-out value during charging will not take effect until the power is switched off and on again. This also includes any changes to the DIP switch settings. Any changes to other settings will take effect immediately and will affect the current charging cycle.

Trickle charge

The trickle charge requirement is calculated by dividing the Ah (amp-hour) rating of the cells by 20. So, for example, if the cells are rated at 2400mAh, then the trickle charge current should be set to 120mA.

Adjusting the dT/dt value

The endpoint temperature rise detection adjustment (dT/dt) should initially be set to 2.5°C per minute (ie, by adjusting VR2 for 2.5V on TP2). In some cases, however, the charger may stop before the battery is fully charged or conversely, it may tend to overcharge the battery.

Under-charging is indicated if the charging period appears to be too short and the batteries do not deliver power for the expected period. In this case, turn VR2 further clockwise to increase the dT/dt value.

Conversely, if the battery pack becomes quite hot after full charge has been reached, turn VR2 anticlockwise to decrease the dT/dt value.

This view shows how thermistor TH2 is attached to a 5.3mm crimp eyelet and fastened to one end of the case.

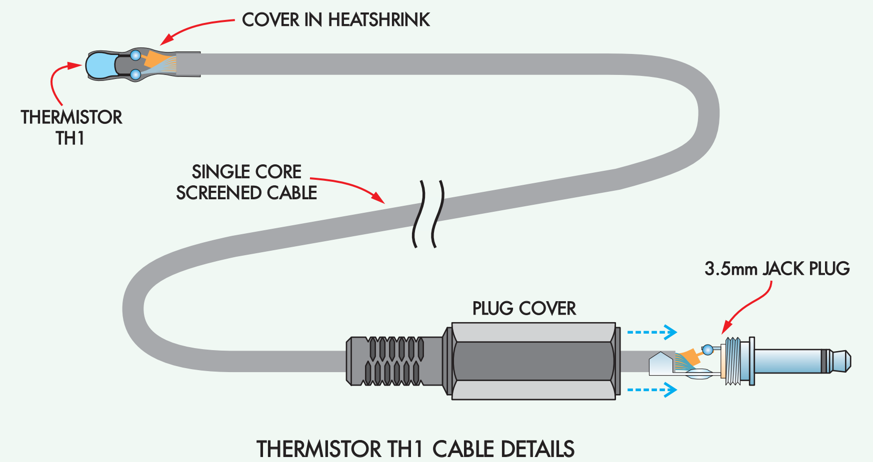


Fig.8: the battery-pack temperature sensor (TH1) is connected to the charger via a length of single-core screened cable and a 3.5mm jack plug. Be sure to heatshrink the thermistor connections so that they cannot short together.

a 3.5mm jack plug via single-core screened cable. Be sure to sleeve the thermistor connections with heatshrink tubing to prevent any shorts between them or to the battery holder terminals.

The thermistor itself needs to be mounted in the battery holder so that it makes contact with the side of at least one of the cells under charge. For our prototype, we drilled a hole in a 2 × AA cell holder so that the thermistor is sandwiched between the cells when they are in place (see photo).

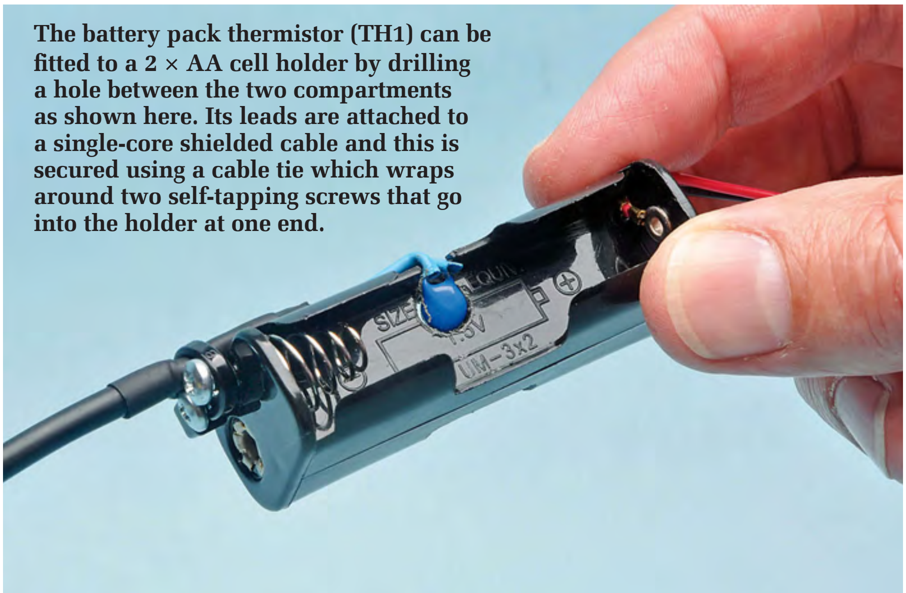
Alternatively, depending on the type of battery holder (or if no holder is used), the thermistor can be held in place against the cells using a length of hook and loop material.

The shielded lead running to the thermistor is secured to the end of the battery holder using a small cable tie and a couple of self-tapping screws.

Setting up

It's now time to make some initial voltage checks. First, with IC1 still out of its socket, connect a DC plugpack to CON1 (positive to the centre of the plug) and switch on. Check that the power LED (LED1) lights, then connect a multimeter between TP5V and TP GND and adjust VR6 for a reading of 5.0V.

Now check that there is 5V between pins 14 and 5 of IC1's socket. If so, check that TP6 is at –5V with respect to TPV+. If this is correct, switch off the power, wait a short time and then insert microcontroller IC1 (notched end to the left).



The battery pack thermistor (TH1) can be fitted to a 2 × AA cell holder by drilling a hole between the two compartments as shown here. Its leads are attached to a single-core shielded cable and this is secured using a cable tie which wraps around two self-tapping screws that go into the holder at one end.

Adjustments

Now for the final adjustments. This involves adjusting the various trimpots for charge rate, cell/battery temperature cut-out, time-out (ie, the maximum time for which the charger operates before it cuts out) and end-point temperature detection. The procedures are as follows:

- **Charge rate:** the charge rate is set using trimpot VR4 and will depend on the mAh rating of the cells or battery. It will also depend on the current rating of the plugpack power supply being used and on the desired charge rate (slow, standard or fast).

Table 2 shows the charge settings for cells/batteries ranging in capacity from 200mAh to 9000mAh. It's just a matter of choosing a charge rate to suit the cells or battery and adjusting VR4 to give the required voltage on TP4.

- **Cell/battery temperature cut-out:** this involves adjusting trimpot VR5 so that the voltage on TP5 is 2.5V when thermistor TH1 is at 25°C. So, if the ambient temperature is 25°C, simply adjust VR5 for 2.5V on TP5.

If the ambient temperature is 20°C, set VR5 for 2.8V on TP5. And if the ambient temperature is 30°C, set VR5 so that TP5 is at 2.2V.

Note that some battery packs will have a thermistor already installed. This should not be used unless it has the same resistance characteristics as the one specified for TH1. It should measure about 10kΩ at 25°C and the resistance should fall with increasing temperature.

- **Time-out:** the time-out is adjusted using VR1. This can be set from 0-25 hours by monitoring the voltage between TP1 and TP GND. The voltage on TP1 directly translates to the time-out in hours, so if it's set to 2.5V, the time-out will be 2.5 hours. And if it's set to its 5V maximum, then the time-out will be 5 hours.

As stated, the No.1 switch in DIP switch S1 acts as a ×5 multiplier for the time-out. So if this switch is set to ON and TP1 is set for +5V, the time-out will be 25 hours. Similarly, if TP1 is set to 1.2V, the time-out will be six hours (5 × 1.2).

The accompanying panel (*Determining The Charger Settings*) tells you how to calculate the time-out value required for the cells used. Table 2 also shows the typical settings for cells of various capacities.

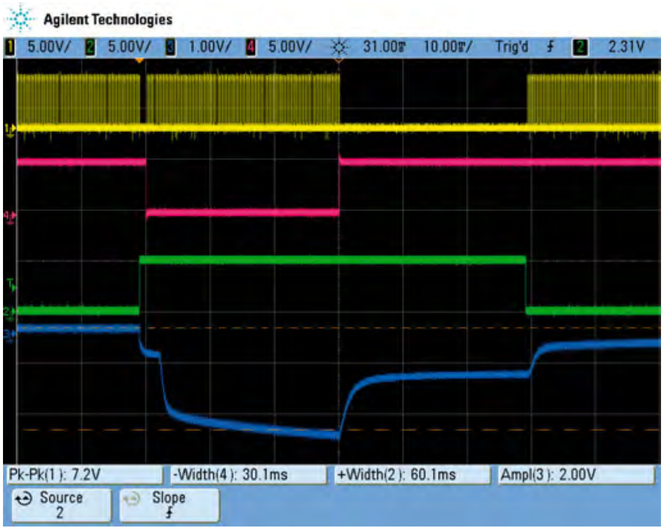


Fig.9: the waveforms in the above-left screen grab show the operation of the *Burp Charger* at a sweep speed of 10ms/div for a 100ms period. The yellow trace is the PWM signal from the microcontroller at pin 6; the pink trace is the 30ms discharge pulse from pin 16 to the base of Q3; and the green trace is the pulse signal from pin 15 to the base of Q4 which turns off MOSFET Q1 while the battery is being discharged and for 30ms after that. The blue trace shows the fluctuation in the battery voltage of a 4-cell Nicad pack. Note that it drops for 30ms (the burp period), then recovers and begins rising again as the charging cycle resumes.

The screen grab to the right shows the operation at a much slower sweep speed of 500ms/div (five-second duration).

Modifications to the prototype

The prototype *Burp Charger for Nicad and NiMH Batteries* included a power switch in the 'Link' position on the circuit (and PCB layout). However, this caused problems because when the switch was in the off position, current could still flow from the external supply into MOSFET Q2 and op amp IC3b via the associated 0.1Ω and 1kΩ resistors. This caused Q2 to switch on and so battery current continued to flow through the 0.1Ω resistor.

This problem was solved by removing the switch and bridging the two switch contacts on the PCB, as shown on Fig.5 (ie, by installing a wire link). The power can now be switched either via the DC plug or at the input power source.

In addition, a 100kΩ resistor has been added between pins 6 and 7 of IC3b. This prevents possible partial conduction of MOSFET Q2 if it has an especially low switch-on threshold. As shown on Fig.5, this 100kΩ resistor is installed under the PCB across the pads of the 100nF capacitor that's also connected between pins 6 and 7 of IC3b.

Note: the power switch was originally installed in the Link position rather than directly after CON1 because we didn't want to burn out the switch contacts by switching the full charging current.

- **Endpoint temperature rise detection:** VR2 is used to adjust the endpoint temperature rise detection (dT/dt). This can be adjusted from between 0.5°C per minute rise to 5°C per minute rise by monitoring the voltage between TP2 and TP GND. Once again, there is a direct correlation between the voltage and the setting.

For example, a setting of 2.5V at TP2 will set the dT/dt value to a 2.5°C per minute rise and this should be the initial setting. Later, this can be changed if you find that the battery pack is either being under-charged or over-charged (see panel).

Top-up/trickle charge options

Setting the No.2 and No.3 switches in DIP switch S1 to ON enables the top-up and trickle charge modes respectively. These can be activated together or individually.

If you want top-up only, set switch No.2 to ON; if you want both top-up and trickle charge, set both No.2 and No.3 to ON; and if you want trickle charge only (without top-up), set switch No.3 to ON (and leave No.2 off).

Note that if either top-up and/or trickle charge is enabled, you then need to set the trickle charge rate (the top-up charge rate is fixed at four times the trickle charge rate). That's done using trimpot VR3, which allows adjustment from 500mA down to less than 20mA.

As before, Table 2 lets you choose the required trickle charge rate to suit your cells. It's then just a matter of monitoring the voltage at TP3 and adjusting VR3 accordingly (eg, 1V = 100mA, 3V = 300mA and 5V = 500mA).

Burp mode

The burp charge mode is enabled by setting switch No.4 in S1 to ON.

Choosing the power supply

As stated, you need to choose a power supply (eg, a plugpack) with an output voltage under load that's at least equal to 1.8 × the number of cells in the battery – eg, 7.2V for a 4-cell (4.8V) battery. Note, however, that the supply must be at least 7V for batteries with less than four cells, to ensure REG1 operates correctly. Refer back to the section titled 'Supply voltage requirements' for the full details.

Half a century of power!

The first 50 years of the 2N3055

by John Ellis

The famous 2N3055 power transistor first appeared in an RCA publication in 1964 (*RCA Transistor Manual*). So, in 2014 it was 50 years old! Few other transistors have survived as long, and it is somewhat ironic that RCA is no longer around to celebrate.

Manufacturing a 2N3055

When RCA developed the 2N3055, germanium transistors were mainstream (see the current series of *Audio Out* in this issue). They were typically made by starting with an n-type 'base' wafer cut into individual chips. A collector was formed by alloying an indium pellet on one side, which overdoped the base to make a p-type layer, and then forming the emitter with a smaller pellet of indium on the other side.

To make their 2N3055, RCA used a similar idea. They started with a thin but whole wafer of silicon, which more conveniently was p-type, and diffused phosphorus into it from both sides to form n-type emitter and collector junctions at the same time. At some point in the diffusion, a base contact region was defined and additional boron introduced on the emitter side to make a low-ohmic base contact. (Further details can be read in RCA's *Solid State Power Circuits – Designer's Handbook SP-52* (1971).) This gave rise to a power transistor with a deep, graded junction. RCA called their NPN process 'hometaxial' because the base region was homogenous (uniform resistivity) in the axial (emitter to collector) direction. It was packaged in a steel TO-3 case.

Robustness

'Second breakdown' is a potentially damaging breakdown condition, which could happen in transistors if a local hotspot forms. This might occur in a part of the transistor that for some reason became slightly hotter than the rest. The base-emitter voltage would then reduce, causing the current to increase further, and thus causing more heating, generating a positive feedback thermal run-away, which usually ends in destruction. Hotspots are prevented by limiting the current a transistor can pass when the collector voltage is high. But, because of the wide base and deep graded junctions, the 2N3055 was robust, and could conduct 1.9A at its rated voltage of 60V (corresponding to full power, 115W) as long as the case temperature was at 25°C or below.

The robustness of the 2N3055 transistor enabled it to be used in linear power supplies and class A as well as class AB power amplifiers. Although it

The original 2N3055 – a hometaxial power transistor made by RCA

also had a maximum current rating of 15A, considerations of gain roll-off and saturation resistance limited the practical maximum current to around 6-8A. Several audio amplifier manufacturers, including famous names such as Quad adopted it for power amplifiers in the region of 50W, making quite a powerful set-up for stereo. It also became very popular with hobbyists and audio enthusiasts.

It is said that imitation is the sincerest form of flattery, and other manufacturers developed similar devices. Mullard, for example, offered a BDY20, and later the BD181-BD184 range; and Siemens offered a BD130. The frequency at which the gain of these devices became unity (the transition frequency f_T) was typically around 1MHz, which was mainly due to the wide base region that was inherent in their construction.

Typical applications

Three circuits illustrate the versatility of the 2N3055. The first, in no particular order, is a linear power supply. When the 2N3055 was announced in the UK it was accompanied with two other devices, the 2N3053, which was a 50MHz, medium current, 40V, planar-type transistor in a TO-5 case, and the 2N3054, a 'baby' 2N3055 rated at 4A, 55V and 25W in a TO-66 case. It was not long before a power supply was published using all three (*RCA Ham Tips*, Vol 27, No. 1, June 1967) A similar circuit is shown in Fig.1, which

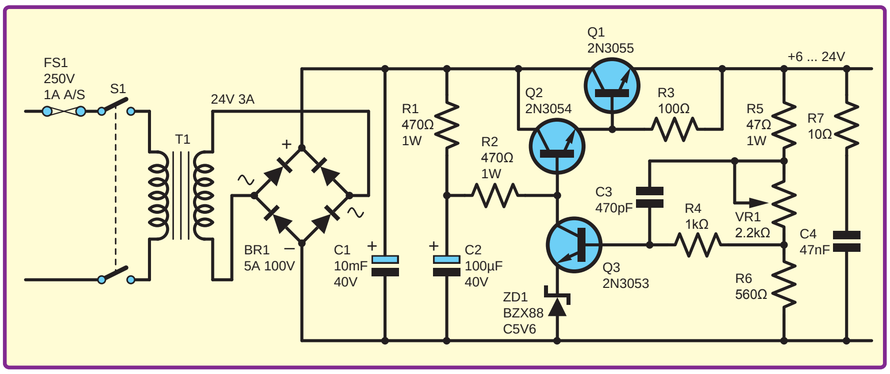


Fig.1. Linear power supply that, in the original design, used a 2N3054 driver, which is now rare, but a BD139 or equivalent device could be used instead, or even an MJE3055T which, in TO-220 packaging is pin compatible with the 2N3054, should anyone need to service any equipment with 2N3054s that need replacing. This PSU will provide a range of voltage output from 6 to 24V. For operation to zero output the reference voltage will need to be taken to a negative supply. Output current is largely limited by the power handling of the 2N3055 pass transistor, which should be mounted on a good heatsink – eg, 1°C/W.

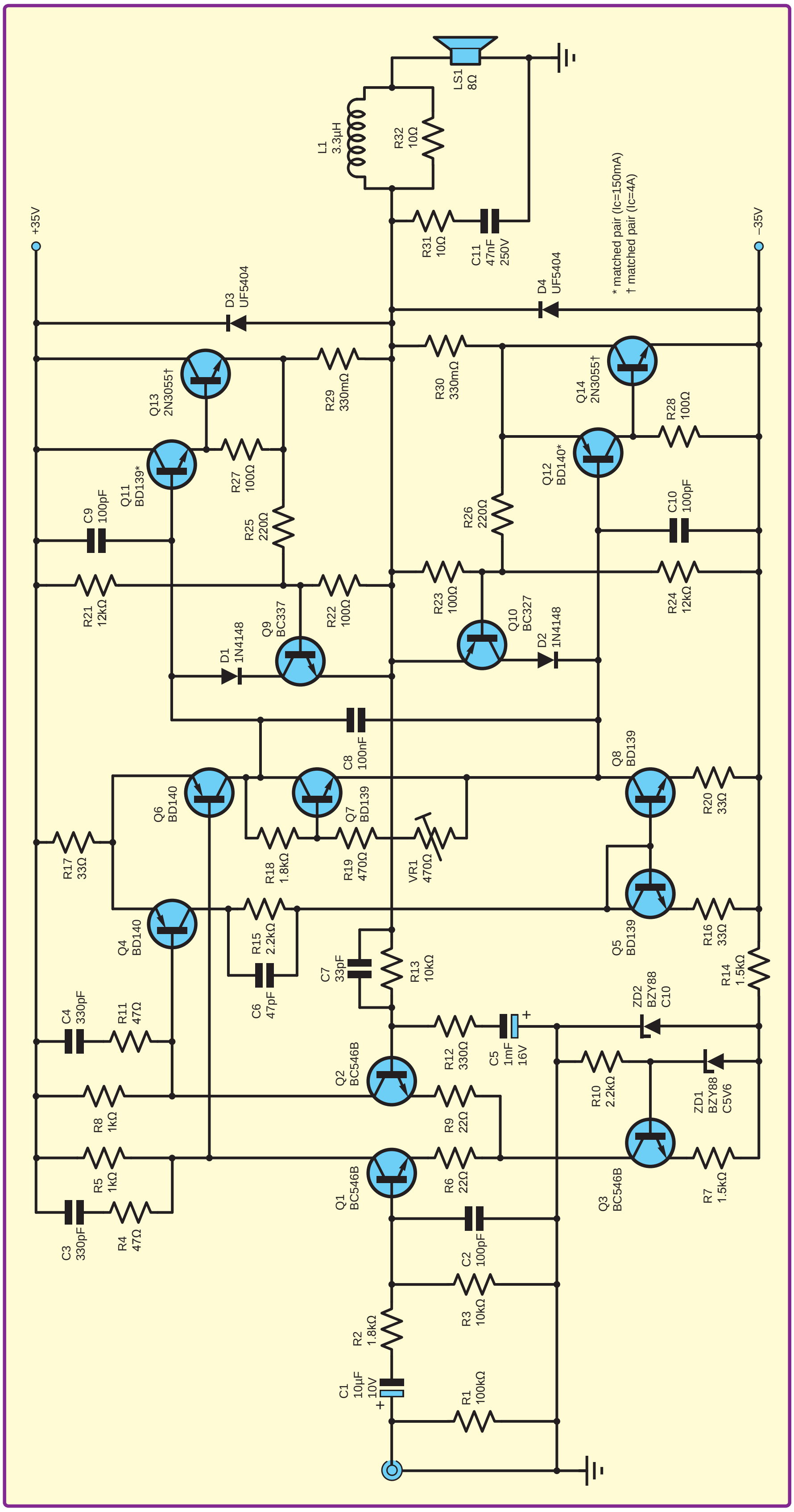


Fig.2. 50W audio amplifier using 2N3055 outputs in quasi-complementary configuration. This design represents an 'intermediate' era between the rather limited performance of standard designs in the 1970s and more modern designs. Distortion is about 0.05% at 20kHz. It uses overall negative feedback frequency compensation and internal 'sprog stoppers' suggested by Ed Cherry as long ago as the 1980s. Unlike normal Miller-compensated designs, in simulation at least, it does not show overloading in the input nor VAS stage for high frequency transients. Interestingly, crossover distortion does not appear to be significantly better using a fully complementary output stage with this feedback arrangement. The differential VAS stage also provides inherent current limiting during overload.

represents the classical 'three-transistor power supply'.

The second circuit is an audio power amplifier. Fig.2 shows a quasi-complementary 50W power amplifier circuit using a pair of 2N3055s, which has low crossover distortion because it does not use classical Miller feedback, and minimal transient distortion (described in the next paragraph). This circuit assumes that the newer epitaxial type would be used (see later), although it could be easily reconfigured for complementary outputs or original-version hometaxial 2N3055s.

The relatively slow frequency response of the original 2N3055 sometimes gave rise to some odd effects in power amplifiers. In 1970, Otala published a paper (Otala, M, *IEEE Transactions on Audio and Electroacoustics*, vol. AU-18, No.3, Sep 1970) describing 'transient intermodulation distortion'. Basically, he pointed out that when an input signal is applied to an amplifier, if the negative feedback were late in arriving, the differential signal across the input stage (whether a singleton or differential pair) could become large and overload the internal stages, causing high but momentary distortion. Many designers recognised this already as 'slew-rate limiting', but there was an even more subtle effect, which may have led to another problem known as 'listener fatigue'. This is where the ear becomes literally tired of listening to an audio signal. One possible explanation is that if the transistor gain reduces (and this could begin at only 10kHz in the original 2N3055) then the only way that negative feedback can compensate is to drive the output stage harder. In turn, the base of the driver transistors need to be fed more current, and many voltage amplifier stages (VAS – eg, Q6 in Fig.2) were biased at around 6mA, which may not have been enough. So this voltage amplifier stage might run out of steam – and clip the current. This would again cause a momentary but high distortion. It may not have been noticed particularly, but the effect would be to detract from the overall quality.

Power converter

The third application area illustrated is in power converters. These are used whenever an AC or DC supply is needed at a different voltage from one which is available. Before transistors, converters used to be rotary – that is, a DC motor driving either an alternator or generator for AC or DC as required, such as used by the army to provide the high tension needed for a valve radio set from the 12V or 24V system available on military vehicles. In the mid 1950s, Royer published a transistor converter circuit which also became popular (Royer, GH, *Transactions of the AIEE, Part 1 – Communications and Electronics*, vol. 74 issue 3, July 1955). In such a circuit, a pair of transistors are operated as switches, driven on and off alternately

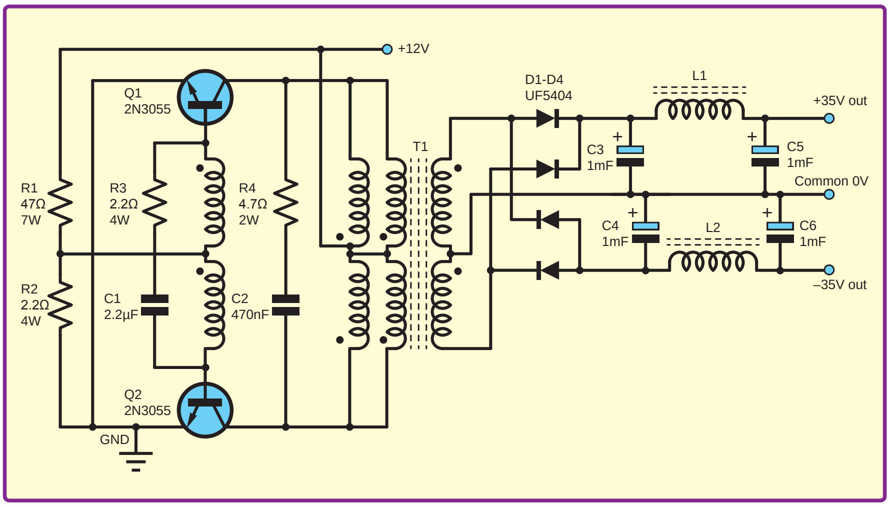


Fig.3. A classical saturated core switching power supply – the transformer is generally not critical and could be made out of an E42/15 core or EDT44 (which are similar, the latter having a round centre core which makes winding thicker wires easier), or an E55 core for higher powers.

Transformer details

Primary – two windings, each bifilar, 2 × 10 turns of 1mm (19 swg) wire on the smaller cores or 1.2mm on the larger, making a total of four windings connected in series-parallel.

Feedback: one bifilar coil, 2× 3 turns, 0.4 to 0.5mm wire (24 to 26 swg)

Secondary: bifilar wound coil consisting of 2× 0.8mm, or 0.9-1mm on the bigger core, approx. three layers giving 30-36 turns each, connected in series to give ±35V output.

The feedback winding needs to be phased correctly to give positive feedback: the windings are all in the same direction start to finish, and the dots then indicate the start ends of the winding. For the secondary, as this is rectified, the only point to check is to connect the two windings in series, if they are in reverse you will get no output.

Typical idling current is 1.5A and with 2N3055s the power output is around 70W at 12V, but can be higher with a 14V supply. Note that heavy power leads should be used to minimise IR drop. The unit should be able to drive at least a 30W amplifier from a 12V supply (or a 50W design to 30W output). Higher powers could be obtained by using parallel 2N3055s or higher current devices – eg, 2N3771s. The output filter (L1, L2, C5 and C6) is recommended if the unit is to operate an audio amplifier to suppress line noise, and the inductors are not critical, about 20 turns on a pot core or RM8, or an RM10-size core with air gap so that it does not saturate at 2A.

These older designs are 'whistlers' because of the 3 to 5kHz oscillation typically used, which is audible. Noise can be largely suppressed by using superglue to join the core halves once the transformer has been checked (there is no way back afterwards!) and also using electrical varnish to suppress wire movement and coil movement on the core; and finally assembling the unit inside an aluminium diecast box.

from a transformer winding from the converter transformer. As transistors could operate at high frequencies, ferrite core designs were used which gave lower losses than traditional iron cores. A typical saturated core converter is shown in Fig.3. Core saturation is used to set the turn-off point, but if a transformer core saturates, the reduced effective inductance allows the collector current to spike just before the transistor cuts out. This increases power dissipation in both the core and transistor, and is now considered to be wasteful of energy, although efficiencies of 80% were achieved. To prevent core saturation, transistors today are often driven from an external circuit (usually an IC) so that timing is set independently from the transformer.

A new process

When RCA developed the 2N3055, it is assumed attempts were made to manufacture a PNP complement. However, no PNP transistor made by a similar process to the hometaxial 2N3055 was ever marketed, so this leads to a conclusion that there were almost certainly difficulties, probably in controlling either boron or aluminium diffusion which would be required to make the p-type emitter and collector regions.

As technology developed, the size of silicon starting wafers increased from the original tiny discs, maybe three-quarters of an inch across, to discs that were one, two then three inches in diameter. One of these transitions proved to be troublesome for the 2N3055, because the wafers became thicker as the diameter increased. The original starting wafers were about 180 microns thick, and the diffusions about 80 microns deep – quite a challenge even by today's standard. Going to a thicker wafer, which is needed to prevent the wafers breaking, meant that the diffusions had to become even deeper, and this proved more or less an insurmountable obstacle. By this time, however, companies like Motorola and Texas Instruments (TI) had developed an epitaxial base power process. In this process, a highly doped starting wafer is used, onto which a lightly doped base is grown epitaxially (meaning that silicon is deposited on the wafer and takes up the original crystal structure) followed by a relatively short emitter diffusion. TI used to call this 'single diffused', seemingly to capitalise on the RCA hometaxial power process, which had an implied ruggedness. Unfortunately, the epitaxial structure gave an abrupt junction rather than a graded one, and as such had a significant second breakdown limitation. This is illustrated in Fig.4 where the safe operating area (SOA) curve is shown for an epitaxial transistor compared with the hometaxial 2N3055.

However, the epitaxial base process had two advantages over the 2N3055. The first was that as the base region was thinner, the frequency response was higher. Typically, this was 4MHz. Second, the process allowed NPN and PNP complements to be manufactured almost as easily as each other. RCA realised that they had to make an epitaxial version of the 2N3055, and after some discussions with JEDEC, the semiconductor registration body in the US, the original device made on the hometaxial process was reclassified in 1978 as the 2N3055H, with the 2N3055 label now referring to an

An epitaxial 2N3055 made by ON Semiconductor

epitaxial device. Motorola, meanwhile, designated a PNP complement they called the MJ2955.

The decline

In 1986, RCA suffered financial losses leading to its acquisition by General Electric (GE). To some extent this was a return to its roots, since GE was partly reponsible for founding RCA in the early 20th century. But RCA was then split up. The hometaxial power transistor range, which included several other popular power devices began to decline as the epitaxial base process was refined. In 2014, ST, a power transistor manufacturer that had licensed the RCA process announced that they were 'obsoleting' the 2N3055. This leaves just ON Semiconductor (formerly Motorola) and a few other smaller manufacturers largely aimed at the replacement market making the device.

So, it seems that the 2N3055 is finally reaching retirement age. The latest audio transistors are out-performing it in terms of gain, linearity and frequency response, and with optimised epitaxial layer and emitter design, they offer higher second-breakdown performance too. The power MOSFET has largely become standard for switching power supplies, and nowadays devices are generally designed for a specific purpose rather than 2N3055's semiconductor jack-of-all-trades role.

Nevertheless, as illustrated by the three circuits shown, the 2N3055 is still popular with experimenters for the same reasons it became popular in the first place – you can use it for amplifiers (even good ones, despite the newer devices), power supplies, converters and power switching circuits which tends to cover most of the circuit applications hobbyists build. The popularity achieved by the 2N3055 is evident by some of the plastic-packaged derivatives that are still available. There is a 75W, TO-220 version, the MJE3055T (which could also replace 2N3054s) and its complement, the MJE2955T; and a 90W variant in TO247 (or originally TO-218) like the TIP3055 and TIP2955. ON Semiconductor also offers a 2N3055H-like device called the 2N3055A.

The 2N3055 might not be around much longer, but it can probably lay claim to having been the most popular power transistor, ever – so, *Happy 50th birthday, 2N3055!*

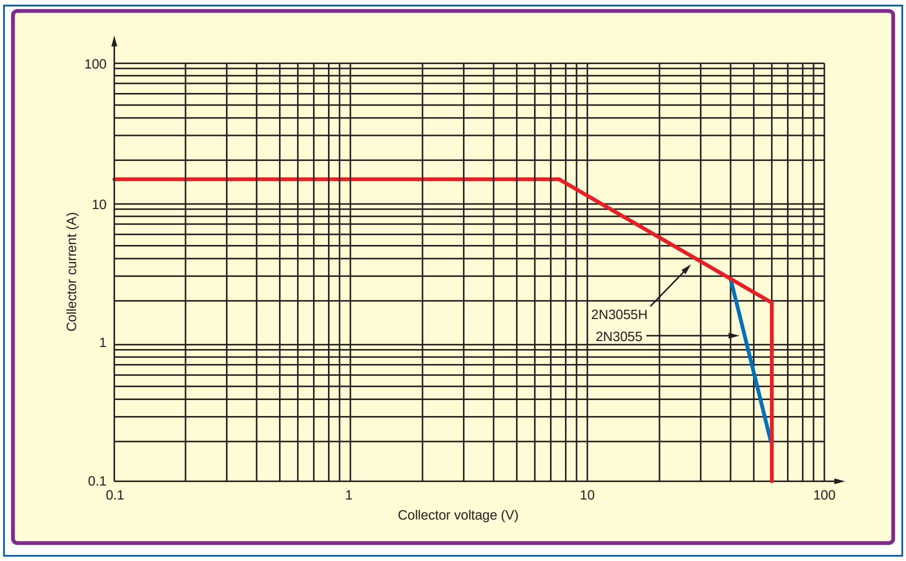


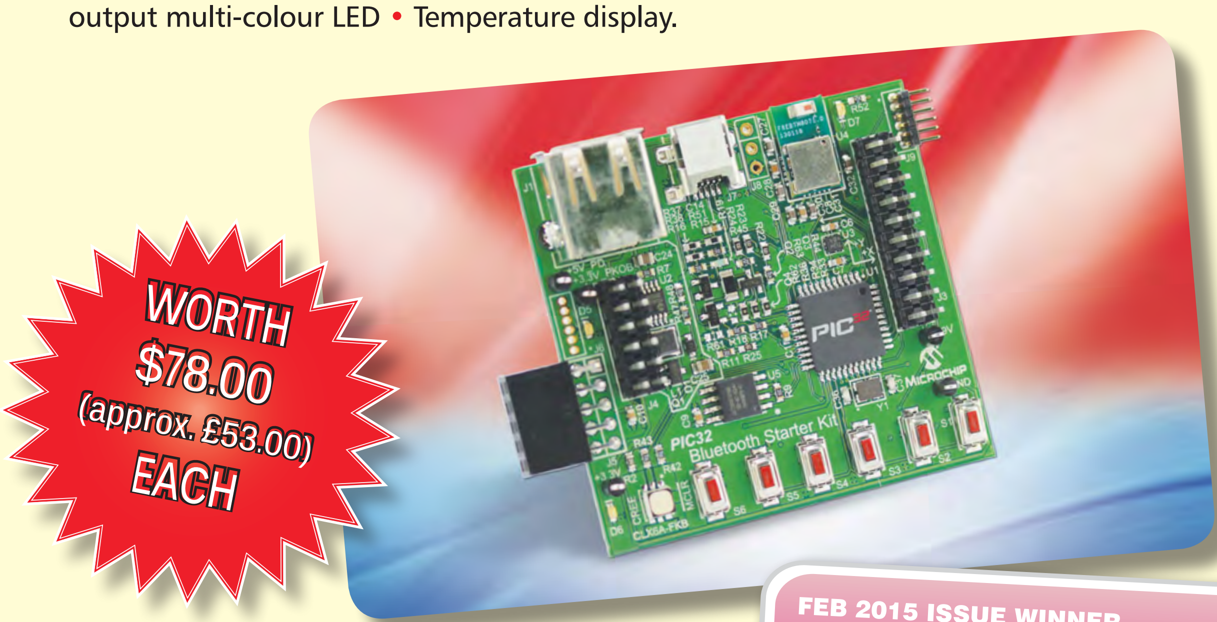
Fig. 4: Graph of safe operating area for hometaxial and epitaxial versions of the 2N3055. It should be noted that ON Semiconductor's latest data sheet gives a better performance for the epitaxial device (0.9A at 60V; nearly as good as the original).

Win a Microchip PIC32 Bluetooth Starter Kit

EVERYDAY PRACTICAL ELECTRONICS is offering its readers the chance to win a Microchip PIC32 Bluetooth Starter Kit. The PIC32 Bluetooth Starter Kit (DM320018) is a low-cost Bluetooth development platform featuring the PIC32MX270F256D MCU. This kit features an HCI-based Bluetooth radio, pushbuttons, Cree high-output multi-colour LED, standard single-colour LEDs, accelerometer, temperature sensor and GPIO for rapid development of Bluetooth Serial Port Profile (SPP), USB and general-purpose applications.

The PIC32 Bluetooth Starter Kit provides a low-cost method for the development and testing of Bluetooth data transfer with PIC32 devices. The starter kit features a Bluetooth Radio, a combination 3-D accelerometer and temperature sensor, 16Mb SPI Flash, an on-board debugger, five user buttons, as well as USB (Device and Host), I²S, I²C, and UART Interfaces.

The Starter Kit offers an Android App, Demo code and Bluetooth Serial Port Profile stack for free to get you started. The PIC32 MCU on the Starter Kit executes Microchip's free Bluetooth SPP stack and Demo Code to perform full-duplex data transmission over the Bluetooth link to showcase: • Colour mixing of Cree high output multi-colour LED • Temperature display.



FEB 2015 ISSUE WINNER
Mr Matthew Wakley, who works at South West Trains, UK.
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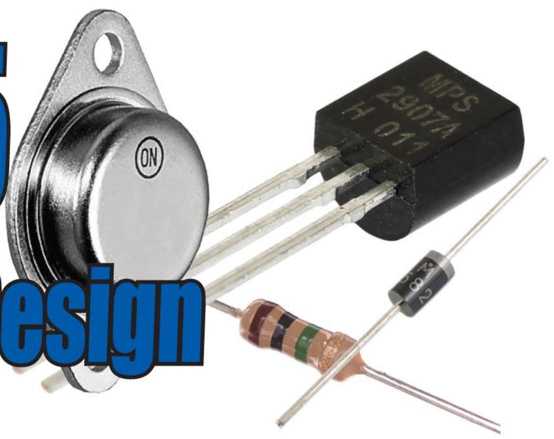
Teach-In 2015

Discrete Linear Circuit Design

Part 5: Filters and tone controls



by Mike and Richard Tooley

Welcome to *Teach-In 2015*. This series is aimed at anyone wishing to develop a detailed understanding of linear discrete semiconductor devices and how they are used in a diverse range of circuits. We hope you will join us on this exciting voyage of discovery! Each part of our *Teach-In 2015* series is devoted to a different aspect of discrete linear circuit design such as modelling and simulation, measurement and testing, noise and distortion. In last month's instalment, *Get Real* explained how we carried out a variety of useful tests and measurements on the *Simple Headphone Amplifier* described in the April 2015 edition of *EPE*. *Knowledge Base* provided you with an introduction to something that we need to avoid in linear circuit design, noise and distortion, while *Discover* described a simple, quick and effective method of checking the frequency response of an amplifier based on square wave testing.

Introduction

In this month's *Teach-In 2015*, *Knowledge Base* introduces the use of filters to modify and/or correct the frequency response of an amplifier or audio system, while *Discover* takes a detailed look at noise and how it can be measured and reduced. Our practical feature, *Get Real*, describes the design and construction of a versatile tone control stage that can be used on its own or in conjunction with the other projects featured in our *Teach-In 2015* series.

Knowledge base: Filters

Simple filters can be built from networks of a few resistors and capacitors, as shown in Fig.5.1. These simple circuits are referred to as 'passive filters' because there is no device present that produces gain, such as a transistor or operational amplifier. Passive filters may, however, be accompanied by some form of amplifier purely for matching purposes, but the active device does not, in such cases, form part of the filter. Active filters are more complex and make use of the gain offered by one or more transistors or operational amplifiers to improve the performance of the filter.

Filters are often also categorised by their order ('first order', 'second order'…). The order of a filter is determined by the number of reactive components (and hence also the complexity) of the filter. It is important to note that the order of a filter has a direct impact on its frequency response; the higher the order the steeper the response will be outside the range in which the filter's response is substantially 'flat'.

Passive CR filters

Two simple passive first-order CR filters are shown in Fig.5.2 and Fig.5.3. Each filter is shown together with its frequency response. The low-pass CR filter shown in Fig.5.2 comprises a series resistance and a shunt capacitor. The reactance of the capacitor decreases with frequency and this has the effect of reducing the amplitude of the output signal as frequency increases (see Fig.5.4).

The high-pass CR filter shown in Fig.5.3 comprises a series capacitance and a shunt resistor. Once again, the reactance of the capacitor decreases with frequency, and this has the effect of increasing the amplitude of the output as the frequency increases, as illustrated in Fig.5.5. Note how, with this arrangement, the amplitude of the output signal increases with frequency.

Cut-off frequency

The term 'cut-off frequency' might imply that there is a frequency within the response of a filter at which the amplitude of the output signal falls rapidly to zero. This is, unfortunately, not the case and the term can, when taken at face value,

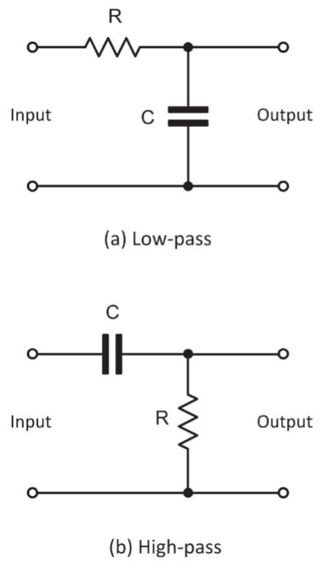


Fig.5.1. Simple CR low-pass and high-pass filters

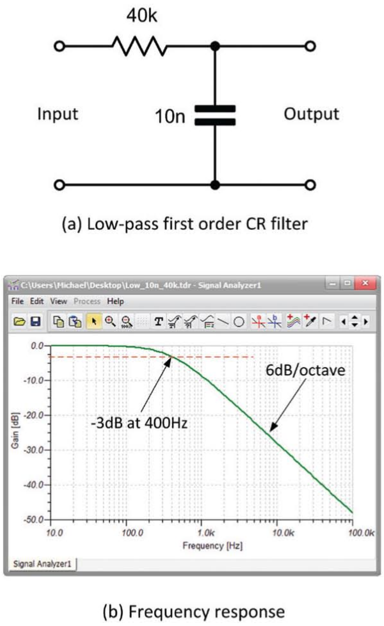
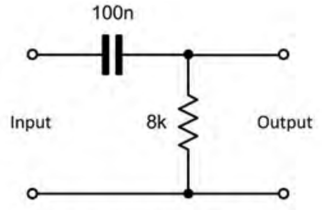
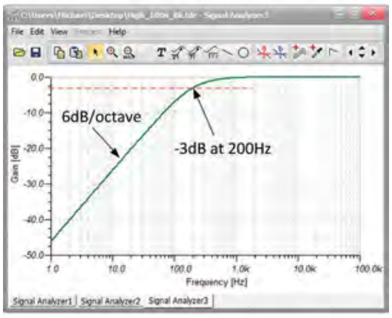


Fig.5.2. Typical first-order passive CR low-pass filter together with its frequency response



(a) High-pass first order CR filter



(b) Frequency response

Fig.5.3. Typical first-order passive CR high-pass filter together with its frequency response

be therefore somewhat misleading. The alternative term 'corner frequency' might be a little more helpful, but it too might imply that there is an abrupt transition between the flat response in the passband and the sloping response in the stopband. In fact, the transition is gradual, as depicted in Fig.5.2 and Fig.5.3.

The values that we have shown in Fig.5.2 and Fig.5.3 are merely representative. To calculate values for a particular cut-off (or corner) frequency we can use the relationship:

$$f_c = \frac{1}{2\pi CR}$$

where f_C is the cut-off frequency (in Hz), C and R are the values of capacitance and resistance (in farads (F) and ohms (Ω), respectively).

At the cut-off frequency, the output voltage will have fallen to 70.7% of the centre of its pass-band value. So, for example, a filter that produces 1V output well within its pass-band will produce only 0.707V at its cut-off frequency.

Practical considerations

To work effectively, our simple first-order CR passive filters need to be driven from a relatively low-impedance source and connected to a relatively high-impedance load. With a resistance in the range 5kΩ to 10kΩ, the authors recommend a source impedance of 600Ω, or less, and a load impedance of at least 50kΩ. This can be achieved by feeding the filter from a simple emitter follower and connecting the filter's output to a second emitter follower (or other high-input impedance buffer stage) as briefly mentioned earlier.

Outside the passband

It's worth carefully looking at the frequency characteristic of the two filters and, in particular, the rate at which the response respectively 'rolls off' above and below the cut-off frequency:

- The first-order low-pass filter (Fig.5.2) was designed for a cut-off frequency of 400Hz. With C = 10nF and R = 40kΩ, this filter starts to roll-off above 100Hz and is 3dB down at the cut-off frequency (400Hz). Above 1kHz (and well into the stop-band) the response falls at 6dB/octave (20dB/decade)
- The first-order high-pass filter (Fig.5.3) was designed for a cut-off frequency of 200Hz. With C = 100nF and R = 8kΩ this filter starts to roll-off below 500Hz and is 3dB down at the cut-off frequency (200Hz). Below 100Hz (and well into the stop-band) the response falls at 6dB/octave (20dB/decade).

As you can see, a particular limitation of these simple first-order CR filters is that they only provide a voltage attenuation of 6dB/octave (20dB/decade) outside the passband. In the case of a first-order low-pass filter this corresponds to a halving of output voltage for every doubling in frequency whereas, for a first-order high-pass filter it corresponds to a halving of voltage for every halving of frequency. These rates of attenuation are insufficient for many applications, but the problem can be easily solved by increasing the order of the filter and making it active rather than passive, as we shall see next.

Second and higher order filters

Simple CR low and high-pass filter networks can be cascaded in order to improve their response. For example, two cascaded CR sections will exhibit a second-order response (12dB/octave or 40dB/decade). Unfortunately, there's a down side to this. Cascading several passive filter stages can have the effect of significantly increasing the attenuation produced by the filter. A better solution is to include one or more active devices in the filter circuit, as described next.

Active filters – the Sallen-Key filter

The Sallen-Key filter is named after its two inventors, who developed the original filter design in 1955. The filter makes use of a unity-gain amplifier that has a very high input impedance coupled with a very low output impedance. An operational amplifier is ideal for this purpose, but an emitter follower will also work reasonably well (in the original circuits a thermionic valve connected as a cathode follower was used as the gain device).

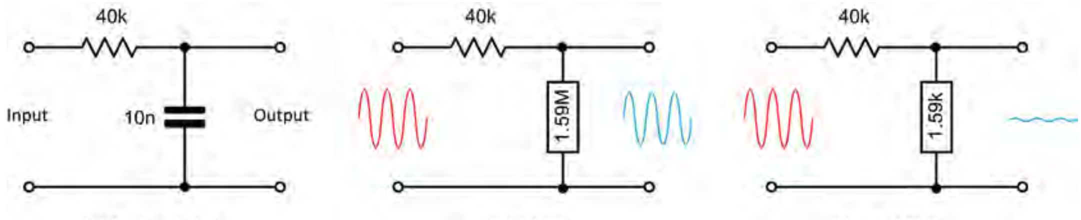
The Sallen-Key filter makes use of two different forms of feedback. Negative feedback (due to the 'follower' configuration, which effectively places the output voltage in series with the input) and positive feedback (through C2 in Fig.5.6 and R3 in Fig.5.7) which is frequency dependant. The result is a second-order response that rolls-off at twice the rate of a comparable first-order filter (ie, 12dB/octave or 40dB per decade). The cut-off (corner) frequency is calculated from:

$$f_c = \frac{1}{2\pi CR}$$

where C = C2 = C3 and R = R3 = R4

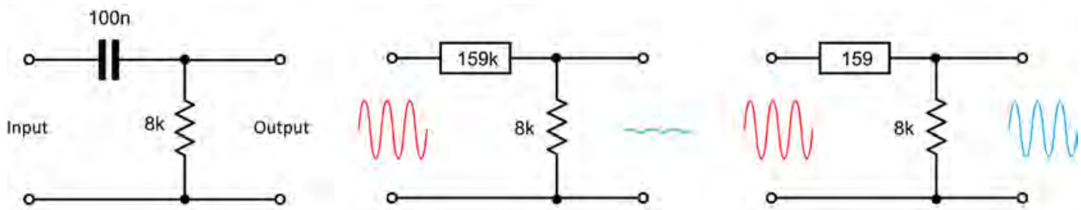
A practical second-order Sallen-Key low-pass filter is shown in Fig.5.6 while its high-pass counterpart is shown in Fig.5.7. In these circuits, the gain device (recall that this must have a high input impedance and a low output impedance for the Sallen-Key filter to work well) is provided by a single transistor (TR1) connected as an emitter-follower.

Base bias for TR1 is produced by the potential divider formed by R1 and R2, along with R5. Note that for predictable operation R5 should have a resistance that is at least ten times the



(a) Typical values (b) At 10Hz (c) At 10kHz

Fig.5.4. Effect of varying reactance on output amplitude in the low-pass filter of Fig.5.2



(a) Typical values (b) At 10Hz (c) At 10kHz

Fig.5.5. Effect of varying reactance on output amplitude in the high-pass filter of Fig.5.3

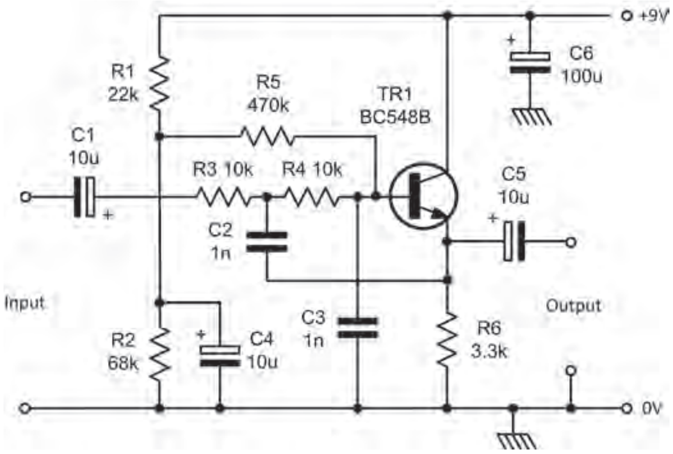


Fig.5.6. Practical second-order Sallen-Key low-pass filter

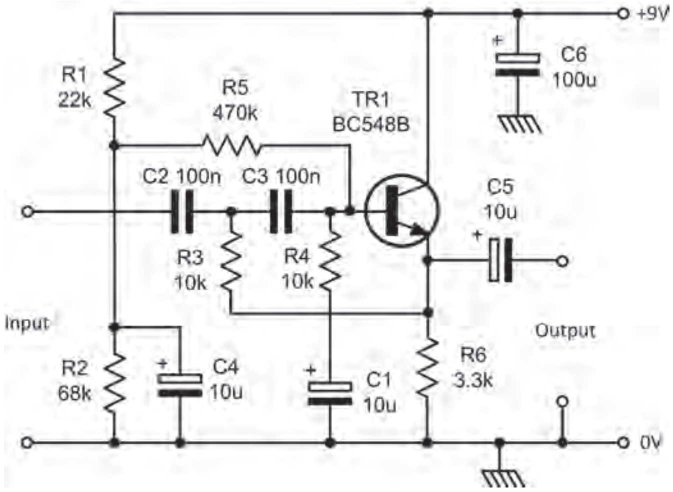


Fig.5.7. Practical second-order Sallen-Key high-pass filter

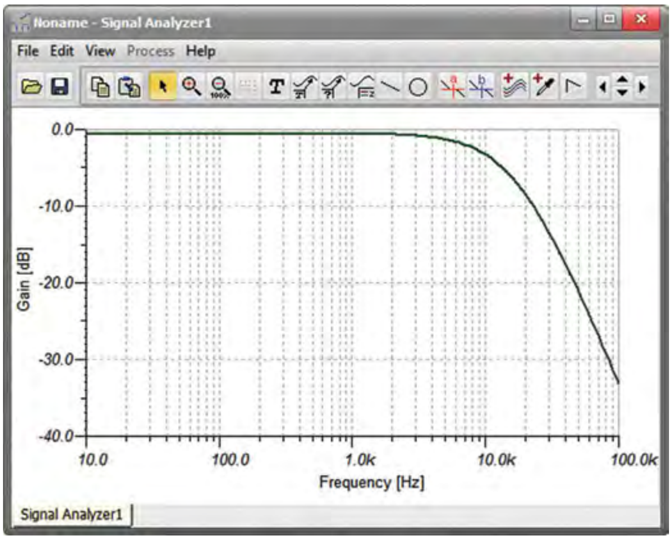


Fig.5.8. Frequency response of the second-order Sallen-Key low-pass filter shown in Fig.5.6

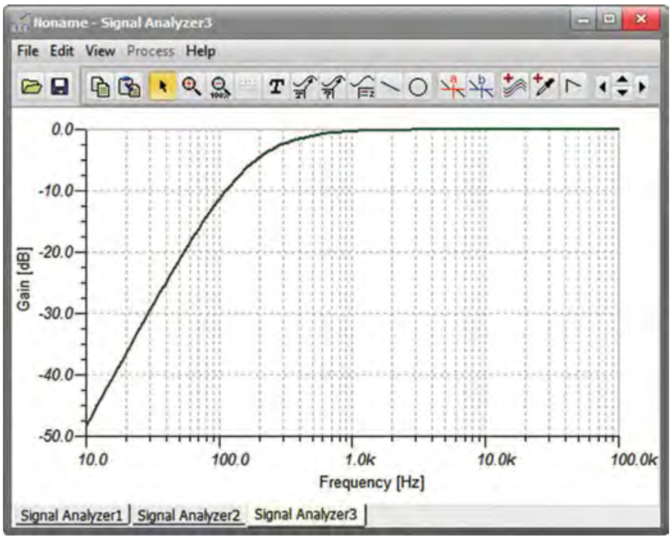


Fig.5.9. Frequency response of the second-order Sallen-Key high-pass filter shown in Fig.5.7

value of the resistors used in the Sallen-Key frequency-conscious network formed by C2, C3, R3 and R4 (recall that, since this is a second-order filter, two sections are used). C1 and C5 provide signal coupling, while C4 and C6 provide decoupling. These two components can also be instrumental in helping to reduce hum and supply-borne noise when the circuit is powered from a mains supply. Within the pass-band the overall voltage gain of the two circuits is unity. The feedback action in these two circuits is sometimes referred to as 'bootstrapping'; a term used to describe the action of a circuit when a change in the output (at the emitter of TR1 in this case) effectively 'pulls-up' the voltage at the input.

The second-order active low-pass and high-pass frequency response characteristics are shown in Fig.5.8 and Fig.5.9 respectively. Note how the filters roll-off at the improved rate of 12dB/octave (40dB/decade) when compared with the simple passive first-order filters. These two filters can be easily modelled using SPICE programs and we have made Tina Design Suite files available for download from the *EPE* website. Fig.5.10 shows the low-pass filter on-test in Tina (note the recommended settings displayed in the Signal Analyzer window). Feel free to experiment with component values (C2, C3, R3, R4) in both circuits in order to obtain different cut-off frequencies.

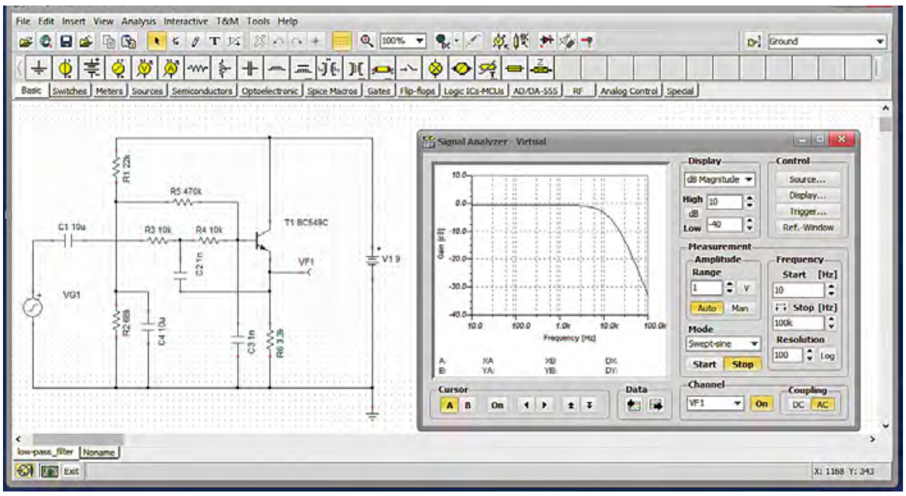


Fig.5.10. Second-order Sallen-Key low-pass filter on test in Tina Design Suite

Discover: Noise

At first sight, noise might not be something that you would actually want to 'discover', but it is something that's present in all analogue circuits and which might become a problem in some situations. As mentioned last month, analogue signals can be susceptible to several different forms of noise including thermally generated noise, shot noise and impulse noise.

Thermal noise

Thermal noise (often referred to as Johnson or Nyquist noise) is caused by the random thermal motion of the charge carriers (normally electrons)

present in all electrical conductors. The implication of this is that every electrical conductor and every resistor produces a small amount of noise. But just how much noise might we expect a humble resistor to produce?

The effective noise voltage generated within a resistor (or any electrical conductor) is given by the equation:

$$v_n = \sqrt{4kTR\Delta f}$$

where v_n is the effective noise voltage (in V), k is Boltzmann's constant (1.372 × 10^{-23} J/K), T is the absolute temperature (in kelvin, K), R is the resistance of the conductor (in Ω) and Δf is the bandwidth in which the noise is measured. Note that thermal noise is essentially white noise, which means that its power is distributed evenly throughout the frequency spectrum.

The formula for noise voltage gives us a few clues as to how we can reduce noise. By reducing the temperature, resistance or bandwidth we can reduce the amount of noise generated. Let's put this into context by seeing just how much noise voltage we could expect to be generated by a 10kΩ resistor in a bandwidth of 50kHz at a fairly average temperature of 20°C (273°C + 20°C = 293K). Inserting the values into the formula that we met before gives:

$$v_n = \sqrt{4kTR\Delta f}$$

$$= \sqrt{4 \times 1.732 \times 10^{-23} \times (273 + 20) \times 10 \times 10^3 \times 50 \times 10^3}$$

$$= 3.18\mu\text{V}$$

Now, 3.18µV might not sound a lot but it is important not to forget that the noise voltage will be amplified along with the signal by all of the subsequent stages. Reducing the temperature to 0°C, the resistance to 600Ω and the bandwidth to 5kHz reduces the noise voltage to a mere 0.24µV. Unfortunately, with semiconductor devices there are several additional sources of noise which we will look at next.

Shot noise

Shot noise results from the passage of electrons across a barrier such as those that exist due to the reverse-biased semiconductor junctions in diodes, transistors and integrated circuits. In this condition, charge carriers (electrons) arrive at discrete intervals rather than as a continuous stream and this can result in a significant amount of noise. Shot noise is often described as similar to the noise of rain falling on a tin roof where, although the rainfall amount might appear to be constant, individual raindrops arrive at discrete times. Note that conductors and resistors do not exhibit shot noise because the current carriers move diffusely through the conductive medium and not as individual 'packets' of current.

Impulse noise

Noise generated by electrical plant (such as transformers, rectifiers, and electrical machines) is, by its very nature, impulsive since it is likely to repeat at the frequency of the supply. Impulse noise can also be generated by static discharge from insulator breakdown and also natural causes such as electrical discharges in the atmosphere. It can be conveyed into equipment and circuitry by various means including stray magnetic and electric fields and along supply rails that are not adequately filtered and decoupled.

Other sources of noise

Other sources of noise (such as flicker noise and burst noise) may also be significant in some electronic circuits. Burst noise, in particular, results from sudden, brief, and relatively large changes in voltage or current. These changes can last for a few milliseconds before returning back to the original level and this can result in recurring popping or crackling sounds in an audio system. Burst noise can be caused by a variety of faults including defective contacts on switches and connectors, poorly made soldered joints, worn or dirty contacts on variable resistors, and manufacturing defects in certain types of electronic components.

Noise and its effect on small signals

We've already shown how thermally generated noise ultimately limits the ability of a system to respond to very small signals. What is of paramount importance here is the ratio of signal power to noise power. Let's illustrate this with an example; suppose the level of the signal applied to an analogue system is progressively reduced while the level of noise generated internally remains constant. A point will eventually be reached where the signal becomes lost in the noise. At this point, the signal will have become masked by the noise to the extent that it is no longer discernible. You might now be tempted to suggest that all we need to do is to increase the amount of amplification (gain) within the system to improve the level of the signal. Unfortunately, this won't work simply because the extra gain will increase the level of the noise as well as the signal. In fact, no amount of amplification will result in an improvement in the ratio of signal power to noise power.

Signal-to-noise ratio

The signal-to-noise ratio in a system is normally expressed in decibels (dB) and is defined as:

$$(S+N)/N = 10\log_{10}\left(\frac{P_{signal}}{P_{noise}}\right)\ \text{dB}$$

In practice, it is difficult to separate the signal present in a system from the noise. If, for example, you measure the output power produced by an amplifier you will actually be measuring the signal power together with any noise that may be present. Hence a more practical measure is the ratio of (signal-plus-noise)-to-noise. Furthermore, provided that the noise power is very much smaller than the signal power, there will not be very much difference between the signal-to-noise-ratio and the ratio of (signal-plus-noise)-to-noise, thus:

$$(S+N)/N = 10\log_{10}\left(\frac{P_{signal+noise}}{P_{noise}}\right)\ \text{dB}$$

It might help to put this into context with some representative figures. Let's assume that, in the absence of a signal the noise power present at the output of a (rather noisy) amplifier is 100µW and when the signal is applied the output power increases to 400mW. The (signal-plus-noise)-to-noise ratio can be calculated from:

$$(S+N)/N = 10\log_{10}\left(\frac{400\text{mW}}{100\mu\text{W}}\right)$$

$$= 10\log_{10}(4000) = 10 \times 3.6 = 36\text{dB}$$

The (signal-plus-noise)-to-noise ratio may also be determined from the voltages produced by an amplifier, in which case:

$$(S+N)/N = 20\log_{10}\left(\frac{V_{signal+noise}}{V_{noise}}\right)\ \text{dB}$$

As a rough guide, Table 5.1 will give you an idea of the effect of different values of signal-to-noise ratio on the quality of a signal when noise is present.

Table 5.1. Effect of signal to noise ratio on signal quality

Signal to noise ratio	Effect (typical)
70dB	Sound produced by a good quality audio system (signal appears completely free from noise)
50dB	Sound produced by a good quality FM radio receiver when receiving a strong local signal (no noise detectable)
30dB	Noise not noticeable – sound quality produced by an AM radio when receiving a strong local signal with no interference
20dB	Noise becomes noticeable – sound quality not sufficient for the reproduction of good quality music but acceptable for speech
10dB	Signal noticeably degraded by noise – quality only sufficient for voice signals
6dB	Very noisy voice communication channel – signal badly degraded by noise
0dB	Noise and signal powers equal – signals severely degraded by noise and of unacceptable quality
–10dB	Signals become lost in noise and unusable

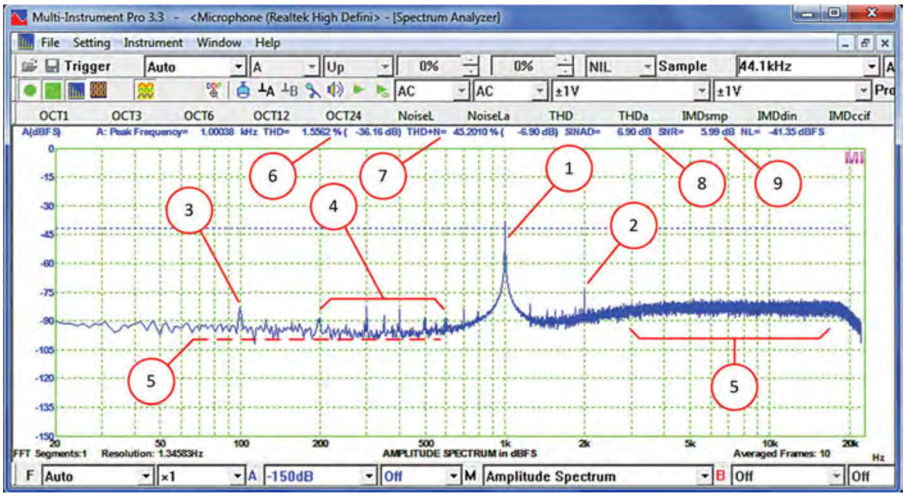


Fig.5.11. Spectral analysis of a signal with noise and distortion

5. A noise floor of about –97dB with a slight increase in noise between about 2kHz and 20kHz
6. A reported THD of 1.5562% (–36.16dB relative to the 1kHz fundamental)
7. A reported THD plus noise (THD+N) of 45.2010% (–6.9dB relative to the 1kHz fundamental) – contrast this with the THD figure without noise!
8. A reported SINAD figure of 6.9dB
9. A reported signal-to-noise ratio of 5.99dB (unacceptably low for most applications).

Hopefully this example has given you an appreciation of just how useful spectral analysis can be when signals are contaminated with both noise and distortion.

Measuring signals in the presence of noise and distortion

As we've already demonstrated in this series, spectral analysis of a signal can be extremely useful in a wide range of practical situations and it can become invaluable when dealing with noise and distortion. As an example, the frequency spectrum of a 1kHz sinewave signal is shown in Fig.5.11 where there is appreciable levels of noise, hum and distortion present (the display was obtained using the Virtins Multi-Instrument that we introduced in Part 1 of *Teach-In 2015*).

In this example, we used a low-level sinewave signal (which had some harmonic content present) mixed with the output of a noise source (see later in this instalment). Take a careful look at Fig.5.11 and note the following:

1. The fundamental of the wanted signal at 1kHz (with a level of about –40dB)
2. The second harmonic of the wanted signal at 2kHz with a level of about –75dB (35dB lower than the fundamental)
3. A component at 100Hz (twice the mains supply frequency) with an amplitude of about –82dB. This was caused by a small amount of ripple present on the DC supply
4. Harmonics of the supply ripple at 200Hz, 300Hz, 400Hz…

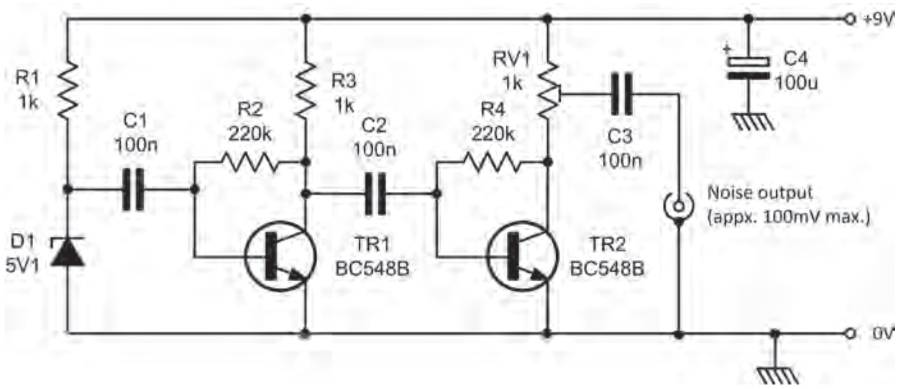


Fig.5.12. The simple noise source

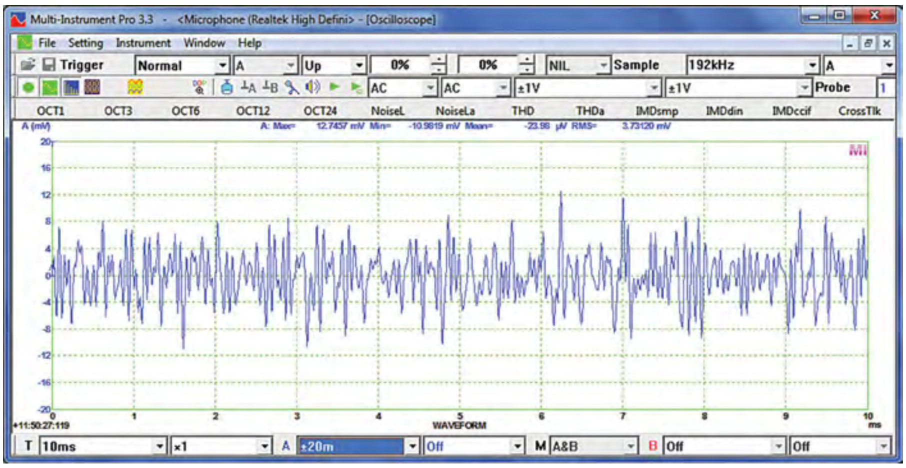


Fig.5.13. Output waveform from the noise source showing an RMS noise voltage of 3.7mV

A practical noise source

A wideband noise source can sometimes be useful when testing audio systems. The source might be described as either 'white' (in which case the power spectrum is evenly distributed) or 'pink' (where the noise output falls at a constant rate of 3dB per octave). A simple noise source can be easily built using nothing more than a handful of common components as shown in Fig.5.12. In this circuit the shot noise (see earlier) produced by the reverse-biased Zener diode (D1) is amplified by the two following wideband common-emitter amplifier stages formed by TR1, TR2 and their associated components.

The output level of the noise source is made adjustable via RV1 up to maximum amplitude of about 100mV. The waveform produced by this circuit is shown in Fig.5.13 – note that this waveform is, by nature, irregular. Construction is straightforward and component values are uncritical so we won't waste space here by describing the circuit in detail. Note that noise sources based on computer sound cards are often severely limited by the upper frequency limit of the sound card. In which case, a 'hardware' noise source is usually much preferred.

Get Real: A simple tone control

Our third *Get Real* project is a tone control that can be used in conjunction with our two previous projects as well as those that follow. The module provides separate treble and bass controls and is based on a circuit that Peter Baxandall developed more than 50 years ago and which, by virtue of its performance and simplicity, has proved immensely popular over the years.

A simplified Baxandall circuit is shown in Fig 5.14. VR1 provides treble control, while VR2 provides bass control (note that we've made a few changes to Peter Baxandall's original circuit in order to simplify the configuration and reduce the number of capacitors used). The 'gain device' shown in Fig.5.14 can take the form of an operational amplifier or one or more transistors.

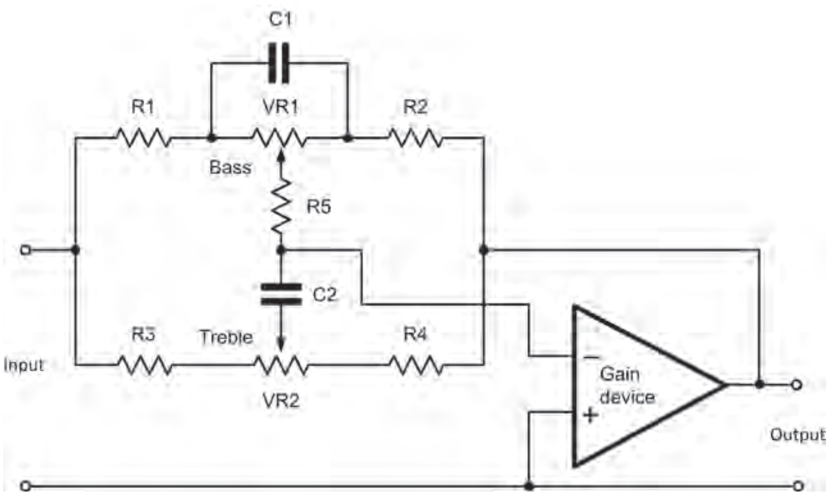


Fig.5.14. Basic Baxandall tone control arrangement

Table 5.2. Outline design specification

Gain	0dB at 1kHz*
Frequency response	–3dB, 10Hz to 100kHz*
Treble control range	> ±12dB at 10kHz
Bass control range	> ±12dB at 100Hz
THD	Less than 0.1% at 1kHz with 100mV RMS input
Max. output	1V RMS (at 0.5% THD) into 10kΩ at 1kHz
Supply voltage	9V at less than 5mA

**Treble and bass controls both set to 'flat' (mid-position)*

Circuit description

Our simple tone control module was developed to satisfy the outline design specification shown in Table 5.2. We wanted to provide a reasonable amount of boost and cut in both the treble and bass frequency ranges (50Hz to 200Hz and 5kHz to 20kHz, respectively).

Circuit description

The complete circuit diagram of the simple tone control is shown in Fig.5.15.

As with the pre-amplifier featured in Part One of *Teach-In 2015*, the circuit uses two transistors, TR1 and TR2. The first stage, comprising TR1 and associated components, operates in common-emitter mode and provides both current and voltage gain. This is followed by the second stage (TR2 and associated components) which operates as an emitter follower, providing appreciable current gain along with a voltage gain of only slightly less than unity.

Base bias for the first stage, TR1, is derived via R8 from the current flowing in the second stage. This negative feedback also helps to stabilise the overall voltage gain and DC operating conditions and allows the circuit to work with NPN general-purpose small-signal transistors. In our simplified Baxandall circuit, VR1 and VR2 provide treble and bass control respectively, while VR3 acts as a volume control. Note that VR1 and VR2 should be linear law potentiometers, while VR3 should have a logarithmic track.

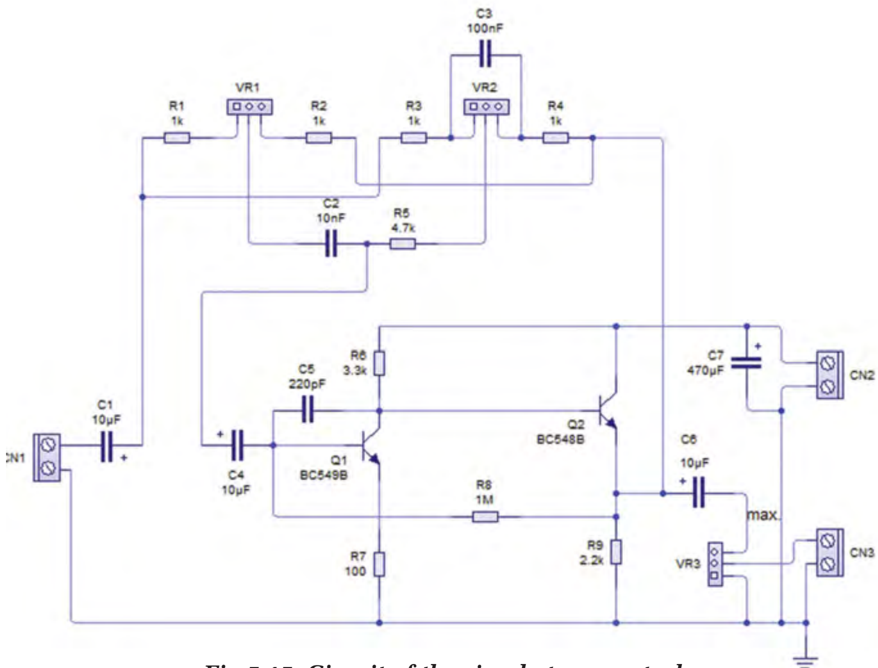


Fig.5.15. Circuit of the simple tone control

Components

General

1 PCB, code 907 available from the *EPE PCB Service*, size 69mm × 48mm
3 PCB mounting 2-way terminal blocks
1 PP3 battery connector
1 SPST on/off switch

Fixed resistors (all are 0.25W 5%)

4 1kΩ (R1, R2, R3 and R4)
1 4.7kΩ (R5)
1 3.3kΩ (R6)
1 100Ω (R7)
1 1MΩ (R8)
1 2.2kΩ (R9)

Variable resistors

2 10kΩ linear potentiometers (VR1, VR2)
1 10kΩ logarithmic potentiometer (VR3)

Capacitors

1 10µF (C1, C4 and C6)
1 10nF (C2)
1 100nF (C3)
1 220pF (C5)
1 470µF (C7)

Semiconductors

2 BC548B (TR1 and TR2, see text)

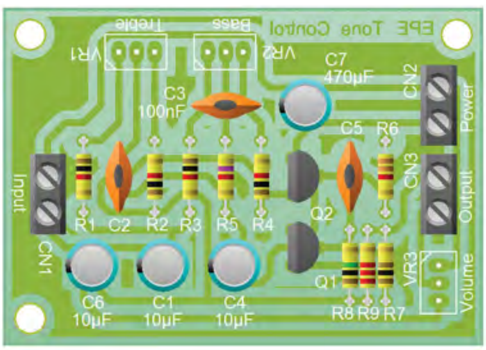


Fig.5.16. PCB component layout shown using Circuit Wizard's 'real world' view

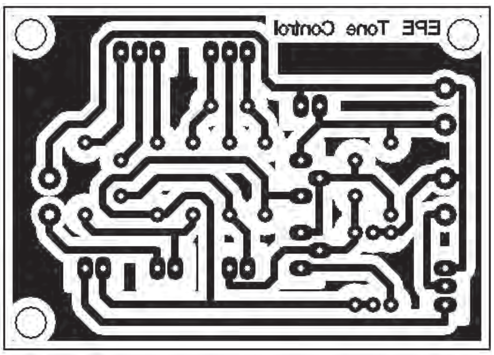


Fig.5.17. PCB track layout

Note that if you are building two tone control boards for stereo operation you will need two sets of components along with dual ganged potentiometers for VR1, VR2 and VR3.

Choice of transistor

We selected low-cost commonly available BC548B transistors for use in the pre-amplifier circuit. They are from the B-gain group of BC548 devices, but any device with a common-emitter current gain (h_{FE}) greater than 150 will

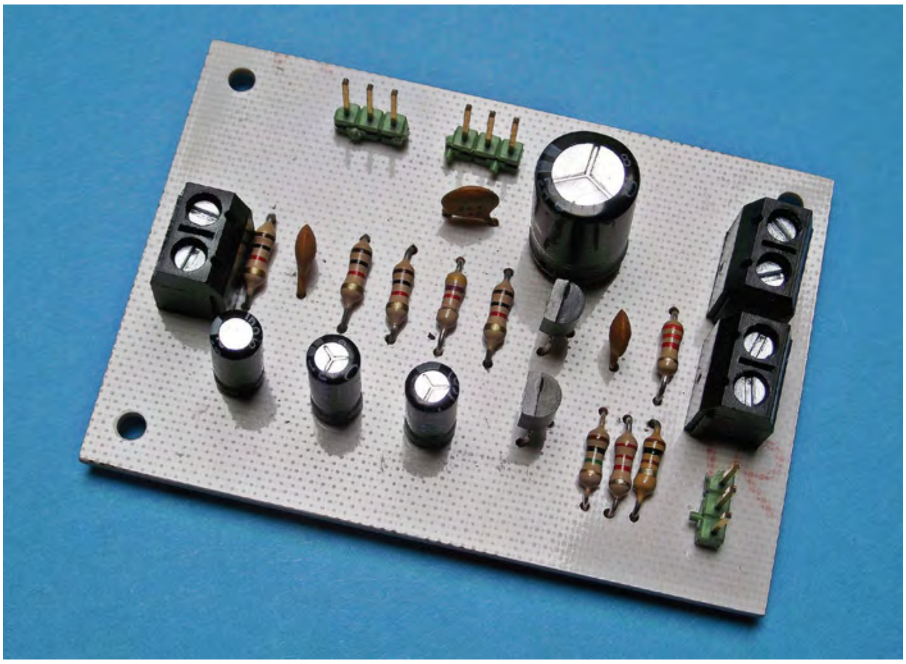


Fig.5.18. Completed prototype tone control ready for testing

prove to be adequate in this application as gain is relatively unimportant. It is therefore possible to use a wide range of general-purpose devices for TR1 and TR2 including BC547, BC549, BC237, BC182, BC167, NTE123AP, and BC550. In all cases it is important to check on the device pin-out before making a substitution.

Construction

Our prototype printed circuit board (PCB) was designed to be built into a small separate enclosure or incorporated into a larger enclosure along with other circuitry and it measures just 69mm × 48mm. As with our other projects, the PCB component layout (Fig.5.16) and copper track layout (Fig.5.17) were produced using Circuit Wizard. The board can be purchased, ready drilled, from *EPE PCB Service*, code 907. The three controls (VR1, VR2 and VR3) are mounted off-board and linked to the board using short lengths of insulated hook-up wire. We used three 3-way 2.54mm pitch PCB headers to neaten our wiring, but this is not essential and the connecting wires can simply be soldered directly to the PCB using the pads provided (see Fig.5.16). Our finished prototype, ready for testing, is shown in Fig.5.18.

Next month

In next month's *Teach-In 2015*, *Get Real* will show you how we used our favourite software applications to check the operation of the simple tone control prior to its construction, and also how we measured the performance of the final prototype design. We will then reveal whether we managed to achieve our original design objectives! *Discover* will be devoted to power and power measurement, while *Knowledge Base* will introduce you to two useful circuit building blocks in the form of constant current and constant voltage sources.

NET WORK

by Alan Winstanley

Certificates of Merit

LAST month's *Net Work* mentioned Superfish adware that found its way by default onto some Lenovo notebook computers. In a technical sense it facilitated a 'man-in-the-middle' attack on the user's supposedly secure web browsing sessions by fooling a web browser into thinking that it was dealing with a remote, secure website; in fact, the session could be hijacked 'in the middle' and images supplied by Superfish could then be displayed in order to 'assist customers with discovering products similar to what they were viewing'. Lenovo then had a *mea culpa* moment and pulled the plug on the idea.

The ordinary web surfer has long recognised secure `https://` websites through the padlock icon that appears in a web browser's address bar. Software such as IBM Security's Trusteer Rappport authenticates known banking secure sites, which ensures users really are dealing with the genuine article and not a phony copy that is trying to steal your logins. (A router's DNS could be hacked to divert users to a pirated copy of a bank website, for example.) A supposedly secure `https://` session encrypts data from end to end like a secure pipeline, giving surfers confidence that the website in question is authentic and no-one is eavesdropping on any activity.

Key to this is the use of a valid secure certificate, signed by a 'Certifying Authority' (CA) which digitally rubber-stamps the holder of the certificate and then provides a digital key (an ASCII file) that is copied onto the webserver's operating system. The host GoDaddy, for example, will sell you a secure certificate for installation onto your GoDaddy website for £40 a year and your domain can then enjoy a golden padlock. Superfish simply worked around all this by certifying its own certificate, so users would think they were dealing with a secure and reputable website.

More recently, another security alert about certificates arose when an Egyptian data company, MCS, was accused by Google of releasing bad secure certificates for several Google-related domains, this time using an intermediate certificate issued by China Internet Network Information Center (CNNIC) which rubber-stamped the certificates with the credentials needed for them to be 'trusted'. The security alert turned out to be down to human error, when an IT system being used by MCS to test a new cloud network under laboratory conditions was accidentally used to surf the web, thereby utilising the highly restricted secure certificate out in the wild. CNNIC, unfairly blamed for the debacle, has revoked its authorisation to MCS and is considering legal action. A similar problem occurred last year when the Indian information agency issued improper secure certificates that could be used for man-in-the-middle attacks or phishing.

Software and web browsers draw on information held in a certificate store – on a Windows system go to Start, search for **certmgr.msc** and press Enter to view them. Firefox often nags users about handling unknown secure websites, especially if accessing a shared web server (one owned by an ISP that is used by multiple clients). The server's secure certificate will therefore belong to the host, not the individual website, which creates a mis-match between domain name and webserver ownership. Although this frequently triggers an annoying Firefox security alert, it's useful to be reminded that the golden padlock system is there for a reason.

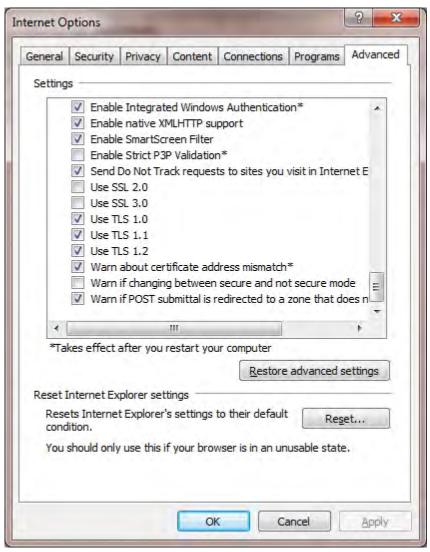
When a secure site is encountered, a chain of events is triggered that causes a browser to check that the relevant secure certificate is still valid. More details of how Windows updates secure certificates is at: **http://support.microsoft.com/en-us/kb/931125**. Note that Firefox stores certificates in its own repository which can be viewed via the Tools/Options tab. Although there isn't much that you can do with certificates, it is useful to know that this process goes on in the background.

Know your SSL from your TLS

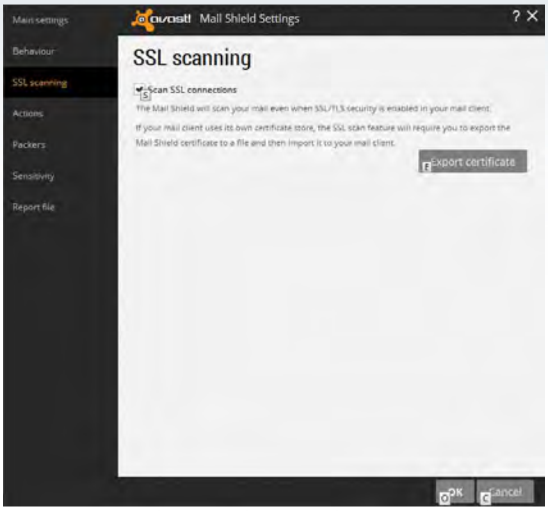
It looks like secure websites might not be so secure after all, with bogus or sloppy secure certificates being issued that can fool web browsers into allowing man-in-the-middle attacks. The risk of users suffering security breaches remains very low though. Apart from web surfing, secure data communications are used by email programs or FTP, for example, and various vulnerabilities in supposedly secure network protocols have also surfaced. Many of them have been created by gaping loopholes in legacy standards that have largely gone unnoticed until now.

Last year, vulnerabilities were discovered in the widely-used OpenSSL that allowed Heartbleed to steal data from a server. This was discussed in July 2014 *Net Work* (also see **http://heartbleed.com**). The 'Secure Sockets Layer' (SSL) is itself relied on by millions of businesses for e-commerce or online security, and it has given us the 's' in '`https://`'. The latest (and last) version 3.0 of SSL had a design problem of its own that under certain circumstances allowed encrypted data to be cracked byte by byte.

SSL was developed by Netscape (famed for Netscape Navigator, the main web browser in use at the time) more



Internet Explorer users should disable SSL altogether and enable TLS1.1 and TLS1.2



Avast Antivirus can scan mail using SSL connections, and also export a certificate if needed

than 20 years ago to facilitate secure web transactions on an emergent World Wide Web. SSL3 is being superceded by an enhanced version TLS (Transport Layer Security). Industry sources suggest that SSL3 is all but obsolete and that it should be disabled in your browser in favour of TLS, to avoid risking any data breach. This is partly because some six months ago another potential man-in-the-middle attack was demonstrated that relied on SSL3; it was found that even if a secure session was started under TLS, if the target server ran SSL3.0 then the browser session could fall back to running SSL3 instead, thereby exposing a vulnerability to data theft. This problem was dubbed 'POODLE' (Padding Oracle On Downgraded Legacy Encryption) or 'Poodlebleed'(!). Security firm Symantec makes it clear that the problem only affects SSL3 running on a server/client and not the secure certificates themselves. More recently, it was discovered that some versions of TLS were susceptible to a POODLE attack as well.

POODLE is another reason to use Firefox – since V34.0 of Firefox SSL3 has been disabled anyway (more details from Mozilla at **http://bit.ly/netwk0614a**). Firefox's deeper settings are accessed by typing `about:config` in the address bar. Users of Internet Explorer should disable SSL altogether in Tools/Internet Options/Advanced... and untick Use SSL2.0 and SSL3.0. Then tick (yes), Use TLS1.1 and TLS1.2 instead.

There will be no substitute for trying some secure surfing sessions and see what difference it makes, but some puzzling SSL software problems do crop up from time to time. My trusted Eudora email program suddenly refused to fetch Gmail from Google, citing a secure certificate error (as Gmail can be fetched more securely via SSL). Avast's Mail Shield had been set to scan email even when SSL was in use. An updated Mail Shield certificate was therefore exported from Avast and imported into the email program's certificate store. If necessary, Avast users can check SSL scanning in: Settings/Active Protection/Mail Shield/Customise/SSL Settings/Export Certificate. Copy the resultant file onto your hard disk and then import it into your email program to update it.

A security freak

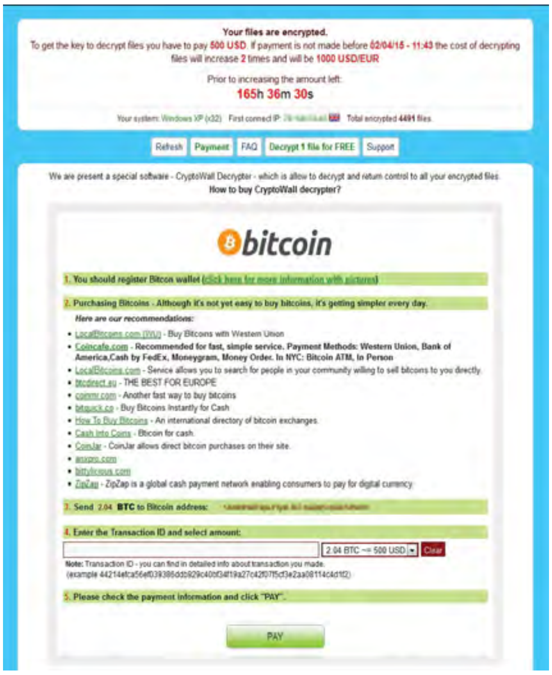
In March this year, security researchers uncovered another poisoned chalice from decades ago, a time when data encryption systems first originated in the US. High-level encryption was considered a weapons-grade commodity, so only a weaker version with a 512-bit key was allowed for export. Over time this lower grade security became embedded in the foundations of emerging software and it is still with us today. Known as 'FREAK' (Factoring attack on RSA-EXPORT Keys), the legacy problem enables the security features of TLS to be bypassed and secure data can be intercepted if accessing a vulnerable website. The problem potentially affects Internet Explorer, Apple's iOS Safari, Opera on Mac and the Android web browser. Thanks to the power of modern cloud computing, the encryption can reportedly be hacked within seven hours, it is claimed, but both Firefox and Google Chrome browsers are reported to be secure. Apple has been tight-lipped except to state that iOS8.2 fixed the FREAK issue, 'by removing support for ephemeral RSA keys'.

The **freakattack.com** website lists numerous websites that are vulnerable. As at 10 March, sites including Groupon, TinyURL, Talktalk.co.uk and IBTimes.co.uk were deemed susceptible. A FREAK test tool is available at **https://freakattack.com/clienttest.html** so you can check whether your own browser is affected; my Android mobile phone failed the test.

It is always wise to be mindful of some of the nastier risks that threaten Internet users. Financial malware such as the exceptionally worrying '**Vawtrak**' Trojan continues to evolve and this software can harvest banking or other logins, disable antivirus programs, disable warning sounds, bypass Trusteer Rapport, log keystrokes and record desktop actions (mousetracks, etc.) as an AVI. It is said to transfer details of the latest command and control servers – the machines at the heart of a botnet – using steganography: the malware uses a favicon (a website's small graphic icon seen in the address bar of most web browsers) to embed imperceptible digital data in the icon and transmit it around the hidden Tor network. Antivirus firm AVG analysed a sample that arrived in a typical spam that we all have probably seen, a fake invoice disguised as a PDF in a phony **.scr** screensaver file. This Vawtrak-infected file was anything but innocuous, and for an insight into the sophistication and criminality behind it, check AVG's report written by Jakub Křoustek at **http://bit.ly/1Ch7DEX** (PDF).

As if to emphasise the critical importance of backing up essential data, a few hours after I started writing this month's column, I was asked to assist a small business that has been hit by a ransomware attack. The firm had fallen victim to **Cryptowall 3.0**, an extremely vicious and menacing malware that heavily encrypts computer files beyond recovery. Over a few days or weeks the PC had become grindingly slow in use, probably because the malware was encrypting nearly 5,000 datafiles, spreadsheets and documents in the background. All documents became unusable and the blackmailers thoughtfully placed some ransom notes on the PC desktop with direct links to a payment page.

The only hope is that, as Cryptowall 3.0 overwrites the original file with an encrypted one, perhaps the earlier version can be recovered in Windows using the shadow copy or Volume Snapshot. See: **http://nabzsoftware.com/types-of-threats/cryptowall-3-0** for some suggestions. Otherwise, unless files can be restored from backups, the only way to recover them is to pay $500 to obtain an unlock key



Cryptowall 3.0 malware encrypts data and blackmails users into paying a ransom in Bitcoins to unlock it

(one hopes – sometimes the key never arrives). A cruel countdown timer shows the time remaining before the ransom doubles to $1,000. The victim is also expected to grapple with Bitcoins, a highly volatile virtual currency that very few can handle. There is no known solution for Cryptowall 3.0 and the firm's infected PC is a total write-off as buying a new machine is a cheaper option.

Thankfully, most of the malware and viruses that are broadcast are intercepted by ISPs and they never make it to our mailboxes. On balance the chances of suffering significant data theft this way are remote; but you never know, it only takes one momentary oversight or mouseclick and some incredibly nasty malware can slip onto your system. Web surfers face similar risks from visiting infected websites, phishing attacks, credit card fraud and identity theft. It has never been more critical to invest in taking robust backups. Macrium Reflect, for example, is an excellent Windows backup program and with bandwidth increasing and cloud storage prices plummeting, also consider hosting backup files offsite as a matter of course. Overall, by remaining very vigilant one can remain secure online, but Firefox and Chrome are the browsers of choice and I always ensure that browsers, plugins and antivirus software are constantly up to date.

FireEye's Cyber Threat Map represents different levels of cyber attacks around the world, 'WarGames style'

The only way to win...

Finally this month, any *Net Work* readers who enjoyed the 1983 movie 'WarGames', which contained the famous line 'The only way to win is not to play!' might check the Cyber Threat Map produced by security firm FireEye (who also offered a free unlocking service for Cryptolocker victims – sadly not suitable for Cryptowall 3 files). It is based on real data and demonstrates how cyber attackers are slugging it out by targeting various countries around the world. A concerted attack (an orange arc) is flagged as an Advanced Persistent Threat (APT). You can check the map at **https://www.fireeye.com/cyber-map/threat-map.html**. Press F11 in your browser for an impressive full screen display.

You can email the author at **alan@epemag.demon.co.uk**

By Robert Penfold

INTERFACE

Level shifting

IN the early days of logic integrated circuits there was a 'Ford choice' when it came to supply voltages – 'anything you wanted, providing it was a +5V supply'. There were some slight discrepancies as to what constituted valid high (logic 1) and low (logic 0) levels, but everything was largely compatible. Things changed when CMOS devices arrived, with their wide supply voltage range and different logic levels. Matters were also complicated by the introduction of various improved 74**** series TTL devices.

Ups and downs

More recently, the main compatibility problem has been caused by the popularity of low-power logic circuits that operate on a 3.3V supply rather than the traditional 5V type. The Raspberry Pi computer is a good example of this, since it operates from a 3.3V supply, and all the GPIO port inputs and outputs work at 3.3V logic levels. A 5V supply is available from the GPIO port, and it will sometimes be necessary to use the port with standard 5V logic circuits. Anyone involved with computer interfacing is likely to encounter plenty of examples of this incompatibility between these two common types of logic circuit.

The best way of dealing with this type of compatibility problem is to simply avoid it as far as possible. It is tempting to use the familiar interface chips that you have used in the past, even if they are not ideal for your latest project, but this is not the best way of going about things. There are now plenty of computer interface chips that operate at the lower supply voltage, or over a range that encompasses both supply levels. With a well-thought-out choice of interface components it is often possible to avoid any supply voltage incompatibility problems.

Simply ignoring the fact that two circuits operate at different levels might actually produce something that works, but this approach cannot be recommended. A 3.3V output might drive a 5V input successfully, but the high logic level from the output could well be in the 'no man's land' between the valid logic input levels of the 5V circuit. If it works at all, results could be unreliable, particularly at high operating speeds.

There is a similar problem with a 5V circuit driving a 3.3V input. The low logic level from the output could be too high, and once again in the 'no man's land' between the two valid logic levels. Also, the high logic level from the 5V circuit is almost certain to be greater than the maximum permissible input voltage of the 3.3V circuit. Being realistic about things, it is highly unlikely that the excessive input voltage would be sufficient to damage the input circuit of the 3.3V device. After all, we are only talking in terms of an overload of about one volt. However, it is quite possible for a slightly excessive input voltage to cause erratic operation, or even a complete malfunction.

Successfully driving a 3.3V input from a 5V output or a 5V input from a 3.3V output is not difficult, and there are logic devices that are designed specifically for this type of thing. For simple applications there is often no need for any special driver chips, and a simple common-emitter switch is all that is needed. A driver circuit of this type is shown in Fig.1. In some cases there is no need for collector load resistor R3, as this function will be provided by a pull-up resistor within the circuit being driven. The Raspberry Pi GPIO inputs, for example, can be set to operate with built-in pull-up resistors.

The circuit can be used with higher voltage logic circuits such as 12V or 15V CMOS types, but the value of R1 should be increased to about 27kΩ. This is necessary because the maximum valid voltage for a low logic level is increased proportionately when CMOS logic devices are powered from higher supply potentials. This type of circuit will also work to provide level shifting upwards, so that a 3.3V logic output can drive a 5V input. A suitably modified version of the circuit appears in Fig.2. The supply voltage on the output side of the circuit has been increased to 5V so as to provide output signals at 5V logic levels, and R1 has been reduced in value to accommodate the lower input voltage range. In other respects the circuit is the same. Bear in mind that a simple level shifter of this type is only suitable for simple applications where high operating speed is not required.

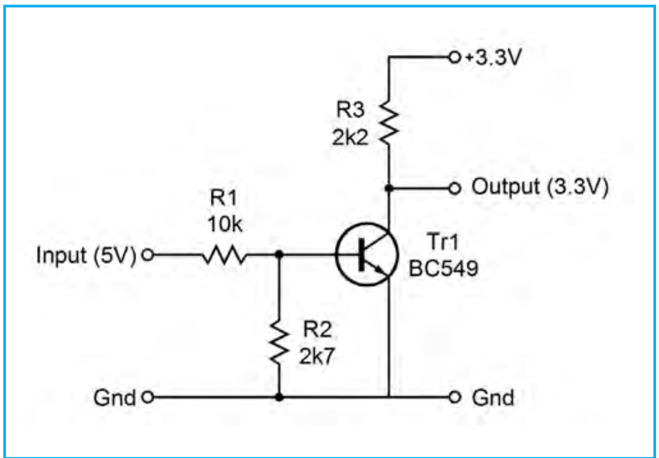


Fig.1. Where high operating speed is not required, a simple common-emitter switch can provide level shifting. Load resistor R3 is not needed if the input being driven has an internal pull-up resistor

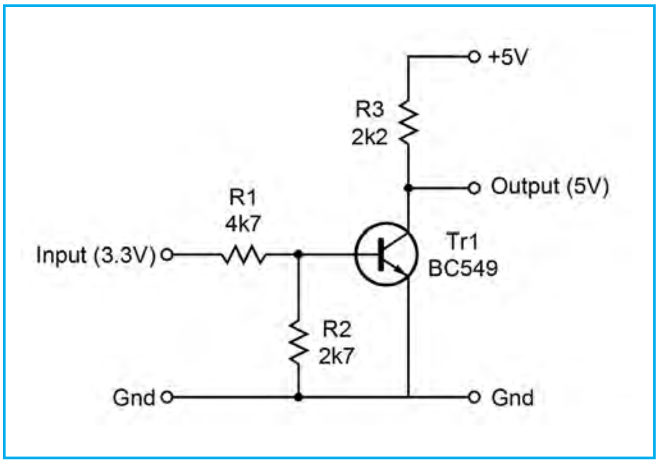


Fig.2. A common-emitter switch can also provide upward level shifting, as in this 3.3V-to-5V example. As before, this circuit is not suitable for high-speed operation

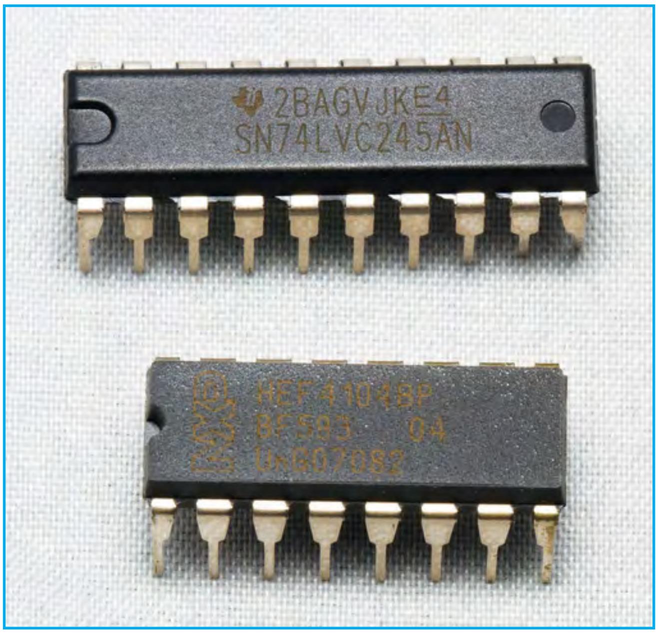


Fig.3. The 74LVC245N chip can provide 8-bit level shifting between 5V logic outputs and 3.3V logic inputs. The 4-bit HEF4104B can provide upward level shifting between logic circuits having operating voltages from 3V to 15V

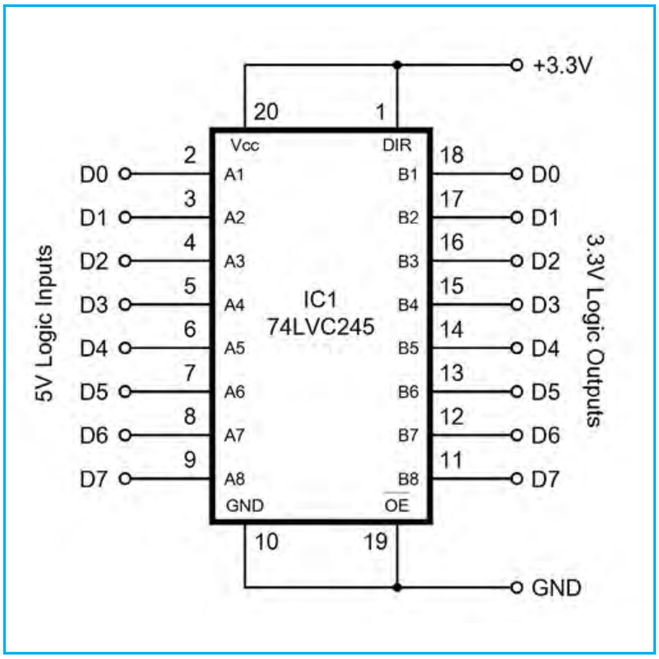


Fig.5. The 74LS245N used to provide 8-bit level shifting from 5V to 3.3V; unused sections should be protected by having their inputs connected to one or other of the supply rails

Downsizing

There are chips that are specifically designed for level shifting in logic circuits, and these are capable of high-speed operation. Two of these, the 74LVC245N and HEF4104BP are shown in Fig.3. The 74LVC245N is a popular choice when driving 3.3V inputs from 5V outputs, and it can handle up to eight inputs/outputs. The original 74245N TTL chip is an octal transceiver that has tri-state outputs. In other words, the eight 'A' pins can be used as inputs with the eight 'B' pins being used as outputs, or vice versa. Either way, the outputs can be switched off and will then go to a high-impedance state. In the current context the bidirectional and tri-state capabilities are of no importance, and it is used as essentially just a normal octal buffer. However, the low-voltage CMOS version – the 74LVC245N – has the useful property of being able to run from a 3.3V supply while still accepting normal 5V logic input levels.

Pinout details for the 74LVC245N are shown in Fig.4 (left). It has a standard 20-pin DIL encapsulation, and conveniently has A1 to A8 and B1 to B8 grouped in order and on opposite sides. The Output Enable input is at pin 19, and this is taken low to enable the device, or high to switch off the outputs. The DIR (direction) input at pin 1 is used to select the direction in which data flows through the device. Data flows from the A bus to the B bus when this input is high, or in the opposite direction when it is low.

Fig.5 shows the circuit for an 8-bit 5V-to-3.3V level shifter using a 75LVC245N. The outputs are permanently enabled by connecting pin 19 to the 0V supply rail, and pin 1 is connected to the 3.3V supply so that the A bus acts as the inputs, and the outputs are available from the B bus. Of course, it is not necessary to use all eight sections of the 74LVC245N. However, this is a CMOS device, and any unused inputs should therefore be connected to one or other of the supply rails. This avoids any build-up of static charge that could cause a malfunction or damage to the chip.

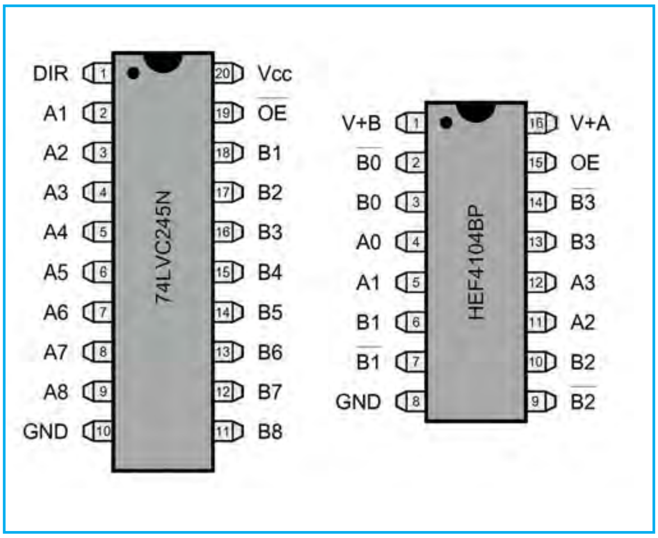


Fig.4. Details of the pin functions for the 74LS245N and HEF4104B chips. These are both CMOS chips and require the standard anti-static handling precautions

On the up

The HEF4104BP is designed for level shifting upwards, so that a low-voltage logic circuit can drive a higher-voltage type. It has a standard 16-pin DIL encapsulation, and its pinout arrangement is detailed in Fig.4 (right). There are separate supply pins for the input and output sides of the device, which are respectively called the 'A' and 'B' sections. The 'B' supply pin is at pin 1, and this must be supplied with a voltage that is equal to or higher than the 'A' supply voltage at pin 16. The HEF4104B cannot be used to provide 'downward' level shifting, and the data sheet makes it clear that trying to use it in this way could damage the chip.

In a Raspberry Pi application, pin 1 would normally be connected to the +5V supply, and pin 16 would be connected to the +3.3V supply rail, so that outputs at normal 5V logic levels would be obtained. This device is not limited to level shifting from 3.3V to 5V, and its recommended supply voltage range is 3V to 15V. It can therefore handle something like a shift from 5V input signals to 10V output levels by simply using these power supply voltages for sections A and B of the device.

There are four inputs, which are called A0 to A3. Their corresponding outputs are B0 to B3, but there are also complimentary 'not' outputs. In Fig.4 these are indicated by the usual bar over the name. In most cases the 'not' outputs will live up to their names and will not be required. The Output Enable (OE) input at pin 15 is taken high to activate all the outputs, or low to switch them to the high impedance 'off' state.

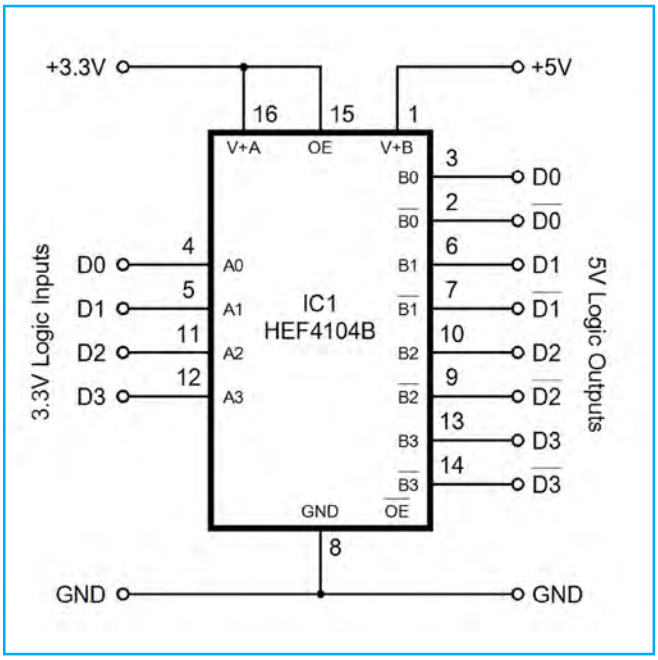


Fig.6. Here the HEF4104B is used to provide 4-bit level shifting from 3.3V to 5V. It can accommodate other supply voltages from 3V to 15V, but it can only be used to provide an upward shift

Fig.6 shows the circuit diagram for a HEF4104BP used to provide 4-bit level shifting from 3.3V to 5V. Of course, it is not necessary to use all four sections, but once again, it is a CMOS chip, and any unused inputs should be connected to ground. As they are CMOS chips, both the HEF4104BP and the 74LVC245N require the normal anti-static handling precautions.

The rest

There are numerous devices intended for level shifting in logic applications, and any Internet search engine should give details of a good selection. Unfortunately, with many of these devices it is much easier to obtain the data than it is to buy the actual devices! Also, while some of these chips are very versatile, they are quite expensive if you can actually find a source of supply. One chip in the 'cheap and cheerful' category that is worthy of consideration is the 74LS07, which is more or less a standard LS TTL hex non-inverting buffer. Being a standard 74LS** chip it is readily available. Details of its pin functions are shown in Fig.7.

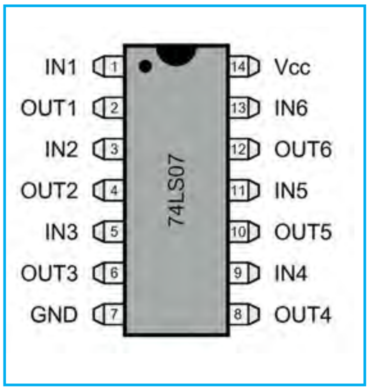


Fig.7. The 74LS07 is a hex buffer, but it has open-collector outputs that can operate at up to 30V. It can only operate from a 5V supply and has inputs that operate at standard LS TTL levels

Being an LS TTL device it operates only from a 5V supply, and it has inputs that operate at normal TTL logic levels. It lacks versatility in this respect, but it is of interest in the current context because it has open-collector output stages. These are used in the same manner as the circuits of Fig.1 and Fig.2, with an external load resistor. The output transistors can be used safely with supply potentials of up to 30V, and they can sink up to 40mA. A shift down to 3V or 3.3V levels can therefore be accommodated, as can a shift up to any normal CMOS levels. Versatility on the input side may be lacking, but the output side can handle any normal logic level.

AUDIO OUT

By Jake Rothman

Ge-mania – Part 3

Fuzzy logic

The most desirable germanium transistors for fuzz boxes are the Mullard OC44 and OC75 because they have the highest H_{fe} values, but they are now very expensive (typically £5 to £25 each on eBay). Another popular one is the Newmarket NKT275, so popular someone has been selling fakes online. Newmarket Transistors Ltd was the germanium manufacturer of last resort in the UK, taking over residual Mullard and Philips production until about 1980. The *Mullard 1979-1980 Data Book* by then listed only nine germanium devices, although Newmarket still lives on as GE Aviation Newmarket. I've got a box of 1000 Newmarket NKT214Fs made in 1972, which makes them relatively recent!

A selection of germanium transistors is shown in Fig.26 (top). The large four-lead transistor TO-7 case on the left is an AF114, which was notorious for internal shorts. Since the shorting whiskers can be fused, transistors can sometimes be 'cleared' by a charged capacitor pulse between the EBC leads (twisted together) and the case/screen lead. Apparently 30mA is sufficient to blow the dendrite whiskers, but they seem to grow again in a few years. Fig.26 (bottom) shows an OC59 hearing-aid transistor which was the first really useful civilian application of the transistor, it's hard to imagine a valve hearing-aid! Also illustrated is a cold-war Soviet Union transistor.

Fig.26 (Top) Popular germanium transistors from left to right: AF114 RF alloy-diffused, OC75 high-gain audio, OC140 NPN switching, NKT214F LF driver and AC188 output; (Bottom) OC45 alloy-junction RF transistor, OC59 hearing-aid transistor, strange Soviet Union device, many on eBay at the moment

Note the old coding system Mullard originally used, such as OC71, based on the valve notation. The C meant a triode and the letter O meant a semiconductor, or valve with no heater. Later, when the Pro-Electron system was introduced, letter A was used to denote a germanium device. Of course B meant silicon and the introduction of the BC107 in 1963 by Mullard heralded the phasing out of germanium. A lot of germanium semiconductors are available using the military CV coding. Examples are CV7005 equivalent to OC71, CV5710 for OC44 and CV7049 for OA10. Most are in *Towers' International Transistor Selector* book. A random plate-full of germanium devices is shown in Fig.27.

Deacy amp

This is the special transformer-coupled germanium guitar amp that apparently gives Brian May of Queen part of his unique guitar sound, used along with a 3kHz 'treble boost' pre-amp. It is based on a circuit from the 1961 *Mullard Reference Manual of Transistor Circuits*. (Note the abstract cover artwork in Fig.28) The story goes that it was built from an old circuit board found in a skip in the early 1970s by bassist John Deacon. It also used a very light paper 6-inch dual-cone speaker. This type of circuit is technically so bad it sounds awful for music reproduction, but amazing on guitar. The distortion is quite complex, there is band-pass filtering, hysteresis and saturation of the output transformer core. Also, the crossover distortion from the class AB output stage is different from silicon; it's softer and less spiky due to germanium's soft, low-voltage turn-on characteristic.

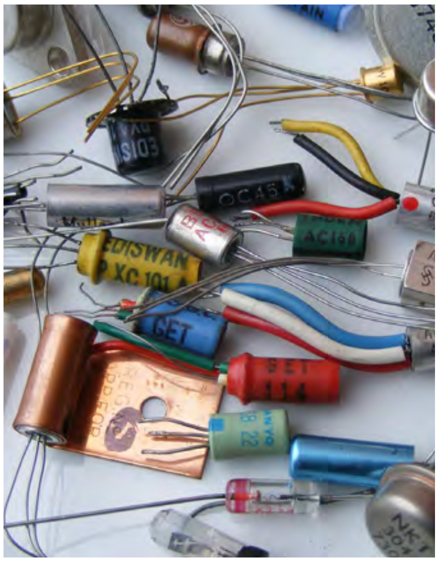
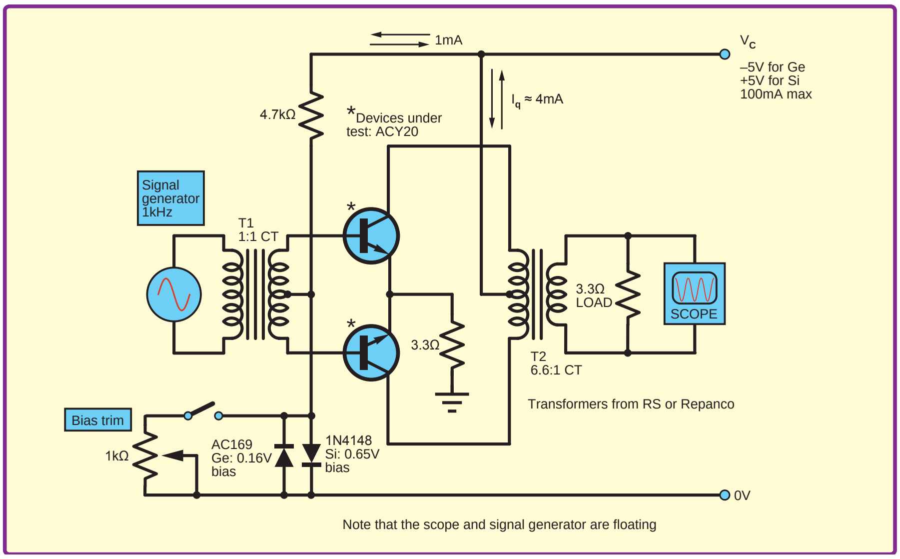


Fig.27. Germanium platter – I miss the colour palette of old devices necessary for human assembly. Today everything is pick and place, and boring black packaging prevails

Note the output transistors must be well matched since it is impossible to apply enough negative feedback to equalise their gains, because of the transformer phase shifts. It is also essential the transistors turn on at the same voltage to minimise the quiescent current needed. H_{fe} matching does not really work, it has to be based on the same V_{be} giving the same collector current. A curve plotter, such as the latest Peak analyser, could be

Fig.28. Old Mullard data books contain very useful information on early transistors and their application. Later books often miss this out, assuming everyone knows. They can often be picked up for 50p at flea markets or charity shops



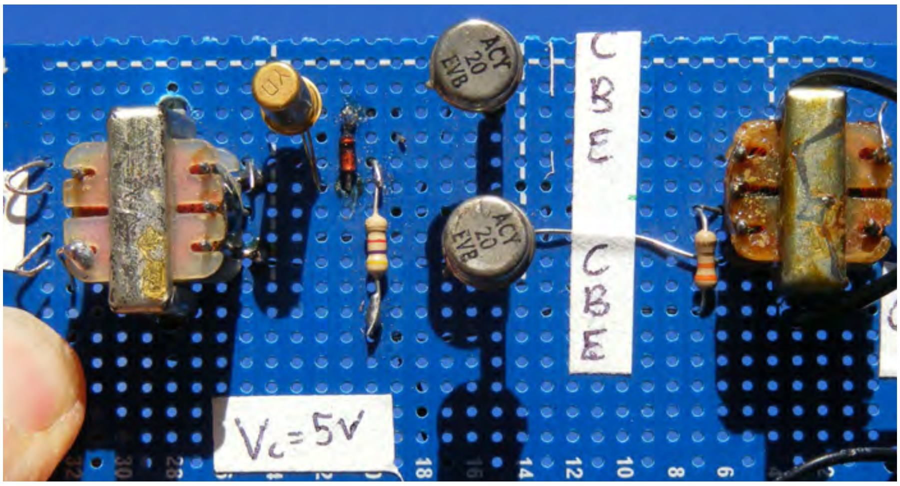


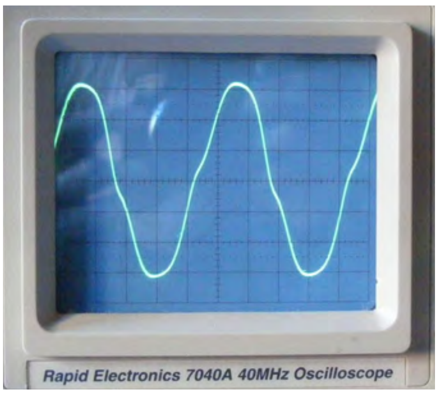
Fig.29 (Top) Circuit for matching pairs of output transistors; (bottom) The matching circuit

used, but when I make one I just match devices using a transformer-coupled output stage with transistor sockets. It runs on 5V to avoid blow-ups. The circuit is shown in Fig.29 (Top) and the board in Fig.29 (Bottom). The distortion builds gradually with germanium transistors, with the crossover bend increasing towards the top of the output range (see Fig.30). This distortion behaviour is very similar to a Marshall push-pull EL34 valve amplifier, except it occurs at one hundredth the power level (*ETI* p.36 Nov 1986). With silicon it is highest at a few mW and is very sharp, more of a switching type effect, see Fig.31.

The common-emitter transformer-loaded output also acts as a voltage-controlled amplifier, since its voltage gain varies with the supply voltage. This gives a compression effect as the battery voltage goes up and down with the signal. To emphasise the effect, put a 56Ω resistor in the power rail with a 100µF decoupling cap, and it pumps smoothly. You can almost emulate this amplifier with a string of effects modules or DSP modelling, but here you can do it with four transistors and it doubles as a 200mW practice amp as well (see Fig.32). If anyone wants to build one (see Fig.33) I've got a large stock of suitable transistors and the Eagle transformers. Surprisingly, new transformers by Xicon are available from Mouser (42TU200-RC), these are half the impedance (200Ω), twice the size and will give 460mW output, see Fig.34. Hammond Manufacturing and Vigortronics still make suitable transformers. The more common LT44 and LT700 can be used for the driver and output transformers respectively, although LT700 will need 12V and a 3Ω speaker. Fig.35 shows an unusual output stage called a 'compound' because it is partially common collector as well as common emitter. It offers lower distortion at low battery voltages and transistor matching is less critical. It was used in the classic Bush TR82 and Ever Ready Sky Leader radios illustrated in Fig.36. In this version, the current in each transistor can be controlled independently.

Germanium transistors have a maximum junction temperature of around 90ºC and are sensitive to soldering, just like LEDs today. This is why many germanium transistors have very long iron leads and should be soldered with leaded solder. They are also easily destroyed by thermal runaway and reverse polarity, and are much more delicate. Strangely, I still see old lecturers telling students to use a heat shunt when soldering modern flow-solderable transistors! Old habits die hard.

I think the Deacy circuit has an educational role in explaining how a push-pull amplifier works and how audio transformers once had a vital role. It also perpetuates the valve amplifier technique of using a phase splitter, which was essential in the era where only one polarity of output device was available.



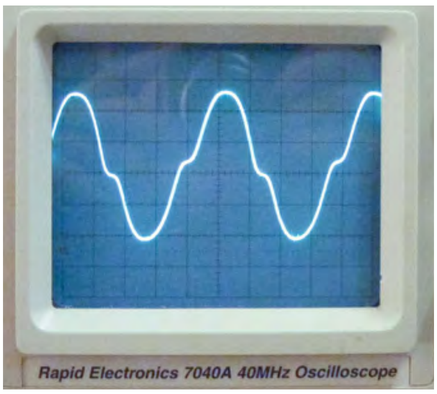


Fig.30. (Left) Deacy amp output distortion is similar to a Marshall amplifier. Fig.31. (Right) Sharper distortion given by silicon transistors in transformer coupled output

Positive earth

Nearly all germanium transistors were PNP, which meant the circuits generally had negative power rails and a positive earth, like some classic cars. This caused confusion for the valve-educated engineers of the time, and still causes short circuits today when germanium fuzz boxes are wired from the same power supply as other pedals. There are some NPN devices available, such as the OC140, which make fantastic fuzz boxes. Interestingly, the OC139/40 transistors are symmetrical, the collector and emitter can be interchanged and the H_{fe} is the

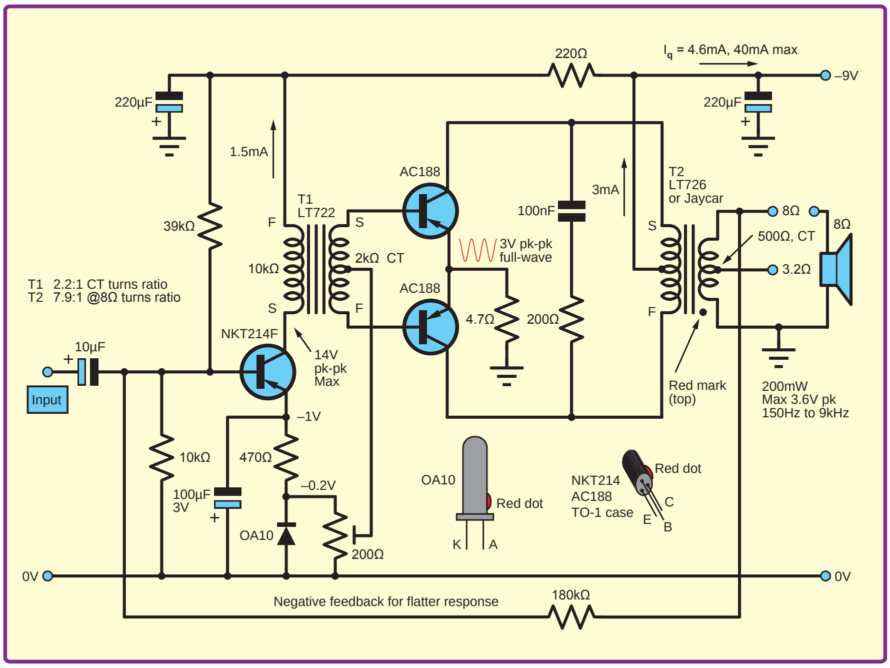


Fig.32. Deacy amplifier, a classic push-pull transformer-coupled germanium design

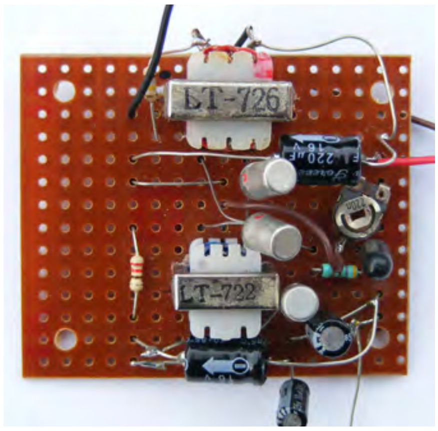


Fig.33. Deacy amplifier built on Veroboard

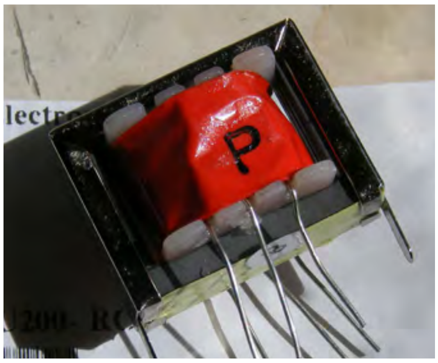


Fig.34. Larger output transformer with lower primary impedance of 200Ω doubles output

same both ways. Most JFETs are similar in that their source and drain can also be swapped over. These properties occasionally give transposed lead-out information with my Peak analyser.

For true complementary power output stages, germanium transistors were initially the only choice. The first complementary output amplifier I could find dating from a March 1960 *Wireless World* used an OC140/OC72 pair and gave around 120mW into an 80Ω speaker. Later, the excellent AD161/2 and the not so good AC127/8 arrived on the scene. The output stage of the EMS VCS3 synthesiser, which is still made today, uses the Siemens AC188K and AC187K. Audio power transistors were the last germanium enclave to survive because they gave a bit more output for a given supply voltage and it was then easier to make matched pairs. Such pairs finally ousted transformers and were very popular in Philips audio equipment, portable radios and radiograms from the mid-sixties to the early-seventies. Before chips, small amplifier sub-assemblies and modules such as the Newmarket PC and Mullard LP series enjoyed about a decade of popularity, see Fig.37. It enabled manufacturers to match their inconsistent transistors together. Mullard and Brimar also did matched transistor packs for their suggested amplifier circuits. The Mullard LFK4 and Brimar LP17 packages were used in many a portable and radiogram. An AM radio based on a Mullard design made by Roberts in 1971, the RIC 2, used a TAD100 chip with a pair of germanium AC187/8 transistors on the output. Such technological combinations were called hybrid circuits and some 1970s televisions would mix chips, valves, germanium, and glass delay-lines on one board like an electronic shantytown.

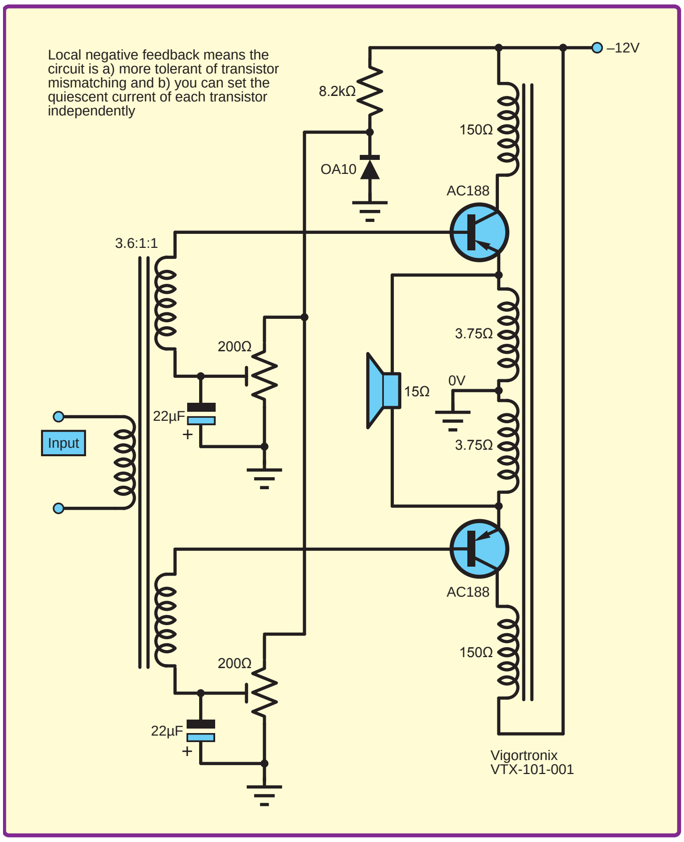


Fig.35. Unusual compound output stage with local emitter feedback gives lower distortion

Sinclair TR750

A small guitar amp using the OC25 is shown in Fig.38 and 39. It uses a FET input to provide high input impedance and hence produce a good top-end (high frequency) response from inductive pick-ups. Germanium transistors are best suited to low

Fig.36. British germanium radios using compound output stages, the small Sky Leader and the industrial design classic Bush TR82

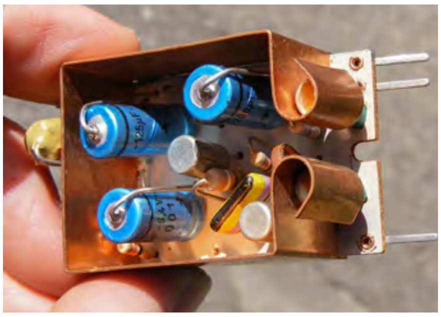


Fig.37. Mullard LP1153 0.5W power amplifier module, roughly equal to a TBA820 chip

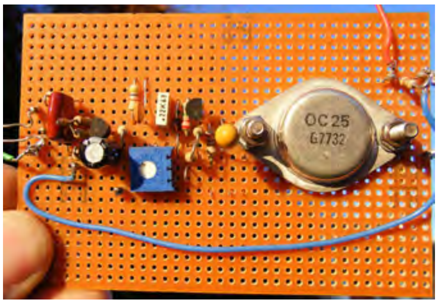


Fig.39. Class-A germanium guitar practice amplifier board

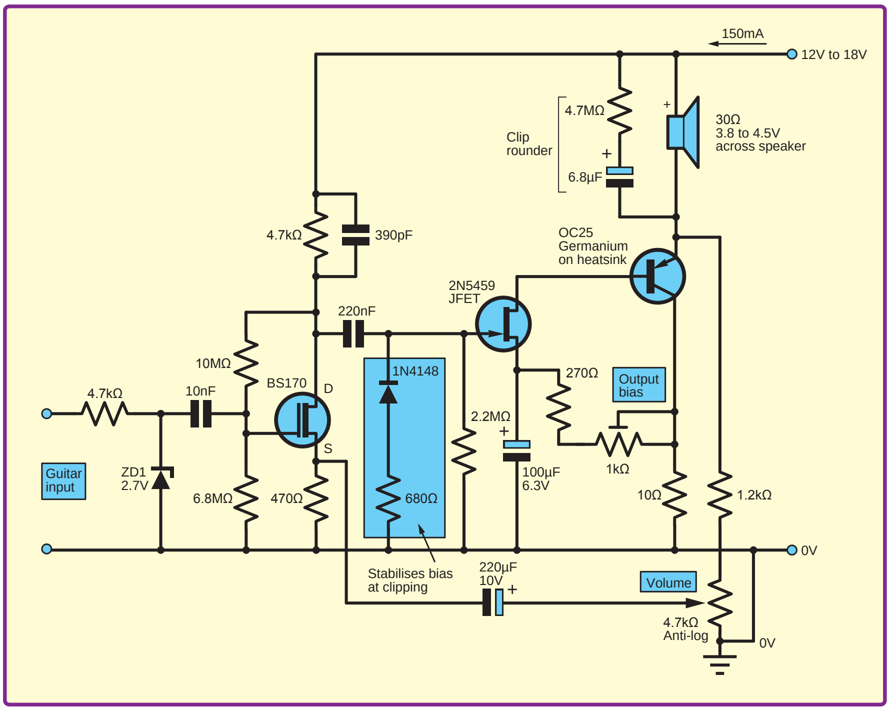


Fig.38. Class-A germanium guitar amp circuit with FET for high input impedance

impedance circuitry. This design was inspired by Clive Sinclair's TR750 amp of 1964. Sinclair bought a lot of double-spread (two-page) adverts in *Practical Electronics* in the Sixties, see Fig.40. The circuit was very simple, as shown in Fig.41; it had no AC feedback and was class A. The original transistors were selected Plessey devices rescued from landfill, a Sinclair speciality and consequently unobtainable through the usual suppliers. I built the circuit and only got 40mW, not the theoretically doubtful, but claimed/specified 750mW. Admittedly I used different transistors, the OC75 and OC25. Suspecting an error in the Sinclair circuit diagram and having no original board to check, I decided to build something of similar simplicity as in Fig.42. I still only got 300mW and it consumed 2.4W, an efficiency of 12.5%, but theoretically correct. I decided to call it the 'Element 32', a semi-tribute to Farnell's latest brand name (Element 14, for silicon of course). (If any readers have experience with the TR750 I would welcome any help or information).

In all these circuits the class-A standing current flows through the speaker, which is an old 30Ω high-impedance 9 by 6-inch elliptical type from EMI (Fig.43) and thus avoiding an expensive output transformer. Because of the DC current and the asymmetry of such circuits, the speaker usually sounds better when connected one way rather than the other! This could be due to the standing magnetic field adding or subtracting from the loudspeaker's magnet or the suspension working better displaced inwards or outwards. In this case, the EMI speaker sounded better going outwards. Such systems had a unique, warm-coloured sound, very easy on the ear for Radio 4 and good for reprocessing samples in the studio.

I do hope you have enjoyed my little detour from silicon down this germanium-paved memory lane. For those who wish to purchase the devices mentioned in these articles, eBay is an easy, if not always cheap route, or you can try me directly via my website: **www.theremin.co.uk**

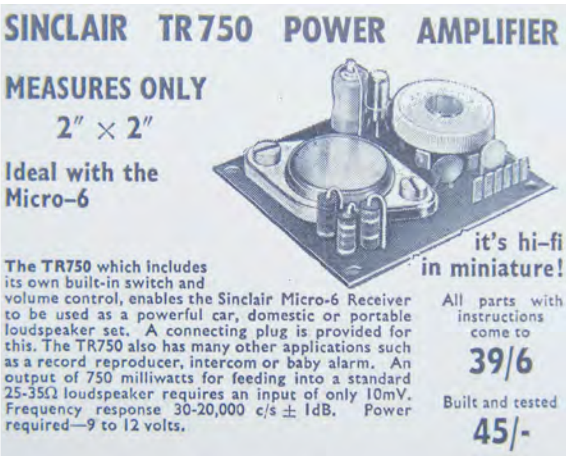


Fig.40. Sinclair advert for the TR750 amplifier module

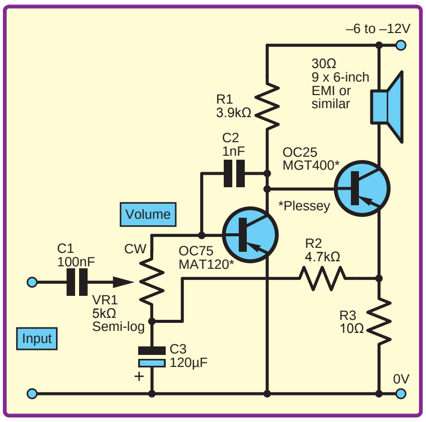


Fig.41. Sinclair TR750 circuit; interesting, but is it correct?

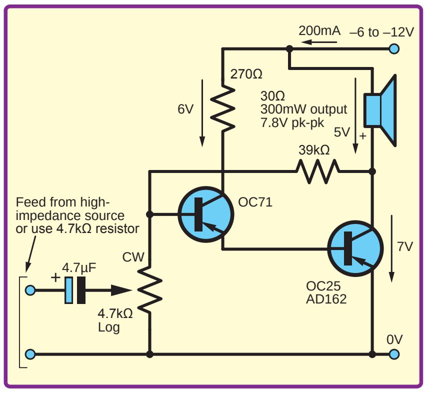


Fig.42. Element 32. My take on the Sinclair TR750, 300mW of class-A germanium power

Fig.43. EMI 30Ω loudspeaker. Many early germanium amplifiers used odd speakers such as this which can be difficult to obtain today

Comparator hysteresis

RECENTLY ***p_j* posted** on the *EPE* Chat Zone to ask for help with a comparator circuit he was working on.

'I am making a room thermostat using an LM34 Fahrenheit temp sensor which outputs 10mV per degree Fahrenheit. I am using an LM324 as a comparator and was looking to control the temperature within 1% with a deadband of maybe 0.5°F or 1°F if the former is not giving the desired results. To achieve this I decided to use hysteresis, but I am having problems as I am using a 5V supply and the output of the temperature sensor at say 72°F is 720mV (fed to inverting input) so the voltage divider I am using to set the required temp at the non-inverting input of the op amp is much larger at the pos end. This means the hysteresis is not well balanced and is virtually non-existent at the lower threshold, ie the temp set at 72°F from the sensor has to rise to over 74°F before switch off but only has to drop to around 71.8°F to turn on. What would be the best way to even this up? I was using a 1MΩ resistor in series with a 1MΩ multi-turn trim pot connected between the op amp output and the non-inverting input to achieve the hysteresis.'

This month we will look at comparator hysteresis, the effect of offsets on comparators and address ***p_j***'s 'symmetrical switching' issue.

It is a common requirement to take the signal from a sensor and determine when that signal crosses a certain threshold level. Typically, this is used to switch a load directly on or off (eg, lighting or heating) or to switch a digital value cleanly between 0 and 1 to indicate 'above' or 'below' the chosen threshold. The circuit used to do this is called an 'analogue comparator' (or often simply a 'comparator').

Comparators used in this way have a fixed reference input to set the threshold to which the varying signal from the sensor or other source is compared. Of course, the reference may be changed in order, for example, to change the temperature at which the load is switched. Comparators can be implemented using op amps (as ***p_j*** is doing) or using one of the many dedicated comparator chips that are available. Comparator chips can provide much better performance, but op amps may be suitable in applications which do not demand high performance.

An op amp circuit without negative feedback has very high gain, thus for all but a small range of input voltage differences the output will be driven hard towards one of the supply rails (that is, the op amp will be saturated). These two voltages (eg, $-V_{sat}$ and $+V_{sat}$) may represent logic 0 and 1 and will indicate which of the two inputs is at the higher voltage (see Fig.1).

As already mentioned, one input of the comparator is usually connected to a fixed reference or threshold level (V_{REF} in Fig.1) and other input is connected to the signal of interest. A real op amp may not switch when the input is at exactly V_{REF} due to offsets. We will look at the effect of offsets in more detail later. Furthermore, for a range of input voltages, the op amp will be in normal 'linear' mode and will output intermediate voltages (see Fig.5 (c)). Although this range of input voltage is very small, in many applications it would be problem if the comparator was 'balanced' between the switching points.

An ideal comparator would switch instantaneously when the input signals crossed the comparison threshold. A real comparator takes a finite period before it reacts; this time

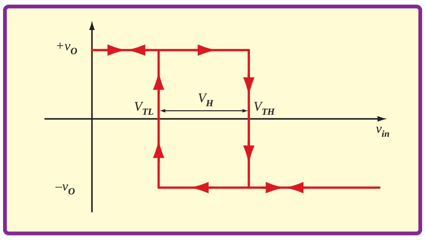


Fig.2. Switching characteristic of comparator with hysteresis

is known as the propagation delay. An ideal comparator's output voltage switches between the two values instantaneously; a real comparator takes a finite time. The rate of change of the comparator's output voltage as it switches is known as its 'slew rate'.

Hysteresis

A potential problem with any comparator is that the output may switch many times as a noisy, slowly changing input crosses the threshold. This is often undesirable, for example if the number of threshold-crossings is to be counted or if rapid switching of the load may be problematical (a light bulb). This may be overcome by using two thresholds, V_{TH} and V_{TL}. The difference between V_{TH} and V_{TL} is called the hysteresis, V_H. When the input increases past the upper threshold, V_{TH}, the comparator switches, but it does not switch back if the input decreases back past V_{TH}. Instead, the input must decrease past a lower threshold, V_{TL}, before the comparator switches again. This is illustrated in Fig.2.

Implementing hysteresis is actually quite straightforward. A comparator has two output states, corresponding to two different output voltages, and these are used to shift the reference voltage. For example, when the comparator is above the upper threshold the reference voltage is shifted down compared to

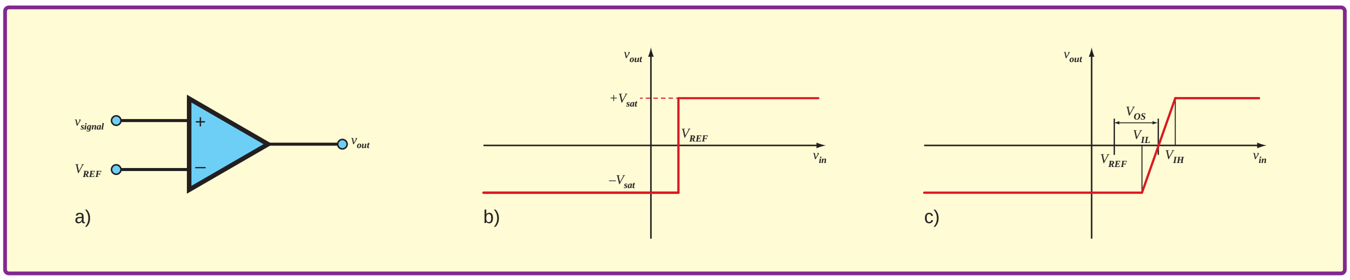


Fig.1. Comparator input-output characteristic, a) schematic, b) ideal response, c) realistic response

the level without hysteresis. This is a form of positive feedback – as the upper threshold is crossed the reference level decreases, so the effective difference between the input and reference increases.

Positive feedback

Thus the positive feedback tends to make the switching action more definite, making it more difficult, or impossible, to balance the comparator between on and off – as could potentially be achieved by inputting a voltage between V_{IL} and V_{IH} for the simple comparator characteristic shown in Fig.1(c). The positive feedback required for hysteresis can be applied to an op amp comparator, as shown in Fig.3. The switching voltage V_{comp} depends on both V_{REF} and V_{out}. V_{REF} will usually be fixed, but V_{out} depends on the current state of the comparator.

In this article we will be using LTSpice simulations to produce illustrative waveforms. The LTSpice schematic used to demonstrate the switching problem just described is shown in Fig.4. Here we have used a generic, rather idealised op amp model, rather than one for a specific op amp. This is because in a later simulation we need to specifically set one of the op amp's parameters to observe the effect it has on the circuit.

The simulation setup shown in Fig.4 uses an LTspice behavioural voltage source B1, to add random voltage variations to the sinewave produced by the standard voltage source V1. Behavioural sources allow the use of a range of mathematical functions to define their output in terms of other parameters in the simulation, such as the current simulation time, and voltages and currents in other parts of the circuit. Further aspects of the simulation set up will be discussed later.

The results from the simulation in Fig.4 are shown in Fig.5. The two plots show the response of the two comparators with the same noisy input signal (green trace V(In_withnoise)). The comparator circuit using U2 in Fig.4 does not have any hysteresis and switches on and off multiple times as the threshold is crossed by the noisy signal (red trace for V(Out_raw)). The comparator circuit using U1 in Fig.4 has hysteresis applied via the feedback resistor R5 and switches cleanly with the same input signal (cyan trace for V(Out_hsyeresis)).

The resistor R6 has little or no effect on circuit operation as the op amp

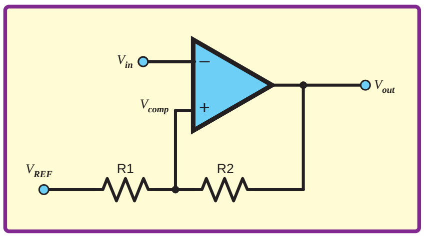


Fig.3. Op amp comparator with positive feedback applied to achieve hysteresis

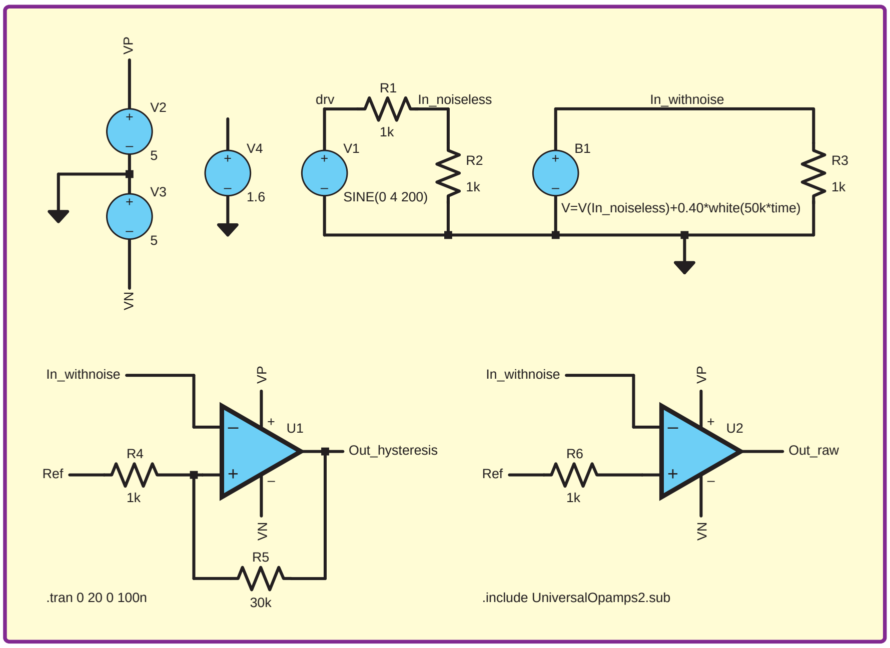


Fig.4. Simulation of comparator with and without hysteresis

(U2) input current is low and B1 is an ideal voltage source. It is included to make the difference between the two circuits minimal. In the circuit for U1, both R4 and R5 play a part in setting the switching levels and hysteresis, and it is useful to know how to calculate these switching points.

Fig.5. The two plots show the response of the two comparators with the same noisy input signal. The upper plot shows a simple comparator switching multiple times as the threshold is crossed. The lower plot shows how hysteresis cleans up the switching

Switching points

Refer to Fig.3 and recall that the switching point V_{comp} depends on V_{REF} and V_{out}. V_{out} can take one of two values; basically the op amp positive and negative saturation voltages, which for simplicity we will assume to have the same magnitude and opposite sign, specifically $\pm V_O$. This assumes a dual-supply circuit (as in the setup in Fig.4). For a single-supply circuit the formulae for the thresholds will be slightly different.

To follow the operation of the circuit, start by assuming that V_{in} is less than V_{comp}, so $V_{out} = +V_O$. As V_{in} is slowly increased this condition remains until $V_{in} = V_{comp} = V_{TH}$ (upper threshold), where:

$$V_{TH} = \left(\frac{R_2}{R_1 + R_2}\right)V_{ref} + \left(\frac{R_1}{R_1 + R_2}\right)V_O$$

This formula is obtained by adding together the individual contributions of V_{REF} and V_{out} to V_{comp}. In both cases we apply the potential divider formula with the other voltage source set to 0V. This is a use of the superposition theorem for electrical circuits that allows the linear circuit to be analysed by considering each source in turn. The overall comparator circuit is not linear, but when we just consider the two resistors with set output and reference voltages the theorem is applicable.

On switching at $V_{comp} = V_{TH}$ the output changes to $V_{out} = -V_O$, changing the threshold to a new value, $V_{comp} = V_{TL}$ (lower threshold), where:

$$V_{TL} = \left(\frac{R_2}{R_1 + R_2}\right)V_{ref} - \left(\frac{R_1}{R_1 + R_2}\right)V_O$$

V_{out} will now stay at $-V_O$ until the input falls below V_{comp} again. The difference in the switching points, ie, the hysteresis V_H, is:

$$V_H = V_{TH} - V_{TL} = \left(\frac{2R_1}{R_1 + R_2}\right)V_O$$

Offsets

The input offset voltage of an op amp used in a comparator circuit will shift the threshold voltages and may lead to incorrect operation. This issue was discussed in ***p_j***'s *Chat Zone* thread. The input offset voltage of an op amp is the voltage, which has to be applied between the inputs to get zero volts at

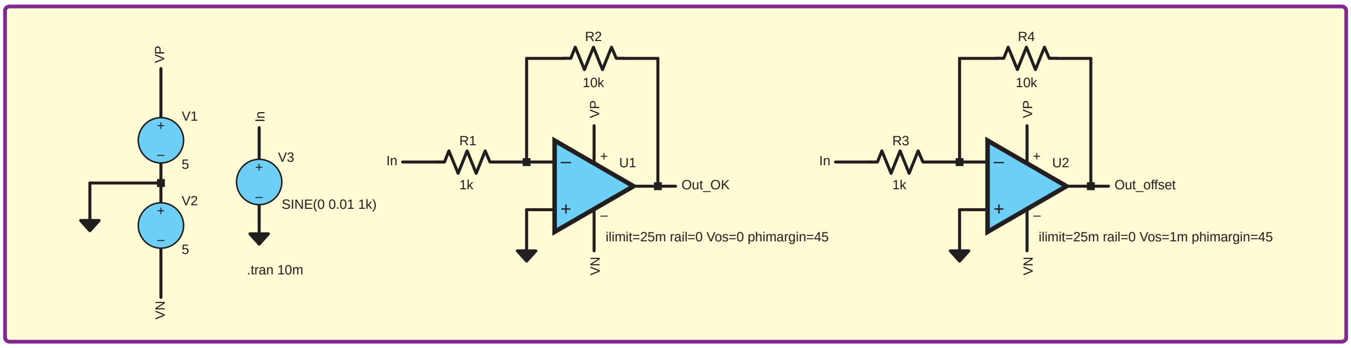


Fig.6. Simulation of an op amp with and without offset

the output. Ideally it should be zero, but for real op amps it is typically in the range of tens of microvolts to several millivolts, depending on the type of op amp used. The LM324 does not have good offset performance – its input offset voltage is in the 2mV to 3mV range.

We can demonstrate the effect of offset with a simulation. As previously noted, we are using a generic op amp model in LTspice. This model can be used by placing the UniversalOpamps2 symbol on the schematic; however, this alone is not sufficient because the symbol itself does not tell LTSpice what to simulate. We also need to include the spice directive on the schematic so that the simulator can find the model file:

```
.include UniversalOpamps2.sub
```

In this case, the model is provided as part of the LTSpice release and is found in the **\lib\sub** folder of the installation; therefore, we do not need to supply a full path. It is also possible to write your own models, or download and use third-party models.

Model parameters

The UniversalOpamps2 model allows us to set a number of parameters (by right clicking on the schematic symbol to open the Component Attribute Editor dialog). We can also choose (using the attribute editor) whether or not each line of parameter information is displayed on the schematic. One of the parameters of the UniversalOpamps2 model is the input offset voltage (Vos), which is what we'll use.

Fig.6 shows a simulation set up where we can check the effect of changing the offset parameter. Here we are simulating a simple inverting amplifier with a gain of 10 (not a comparator). We have two copies of the circuit, the circuit using U1 for which Vos = 0 (the ideal case) and the circuit using U2 for which Vos = 1m (1mV).

Fig.7 show the results of this simulation. We see that for the circuit with offset (red trace V(out_offset)) the op amp's output is shifted up by a constant DC value of about 11mV. This shift is equal to the gain of the circuit with respect to the offset times the input offset voltage (11 × 1mV, 11 is the noise gain, equal to the non-inverting gain).

Having established the approach to setting offset voltage in the op amp model, we can apply it to a comparator circuit. Fig.8 shows the setup used. Here we have a comparator with V_{REF} = 100mV and V_{out} = ±5V, depending on the switch sate. Using the above equations we get V_{TH} = 104.9mV and V_{TL} = 94.9mV. As previously, the simulation uses two copies of the circuit, one without offset (using U1) and one with (using U2). The offset for U2 is 10mV.

The results of the simulation are shown in Fig.9, where we can clearly see that the two comparators do not switch at the same points on the input waveform. Fig.10 shows a zoom in on Fig.9 so that we can see the switching voltage more clearly. Here it can be seen that both thresholds of the comparator using the op amp with offset have shifted by the offset voltage – they are 10mV higher in both cases (red waveform V(out_offset)), thus the comparator in this case switches later on a rising input and earlier on a falling input.

The effect of offset in comparator circuits can be very significant; in ***p_j***'s case a 3mV offset corresponds to a 0.3°F error, which is far from trivial if the aim is to achieve about 1% accuracy. However, it seems this is not the main issue which ***p_j*** is struggling with in his question. As far as we can tell from the question, ***p_j*** is assuming that the reference voltage will always be at the midpoint of the hysteresis, so that we would have $V_{TH} = V_{REF} + V_H/2$ and $V_{TL} = V_{REF} - V_H/2$. Unfortunately this is not the case, which is what we think ***p_j*** means when he says the hysteresis is 'unbalanced'.

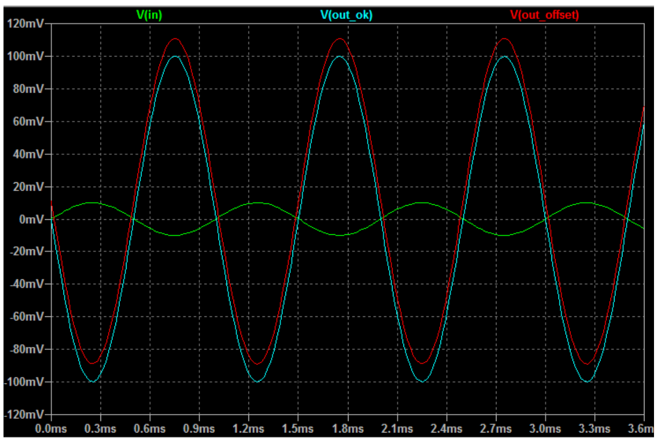


Fig.7. Simulation of effect of input offset voltage on an op amp amplifier

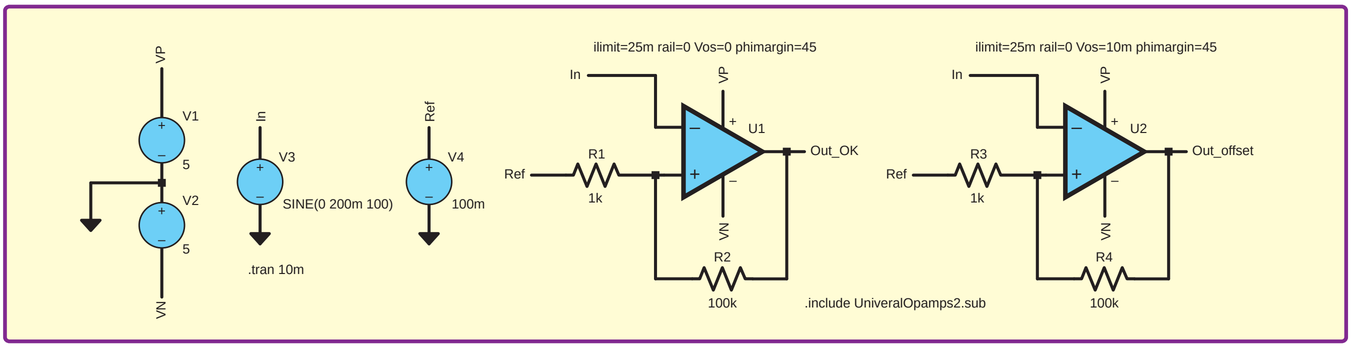


Fig.8. Simulation of comparator with and without op amp input offset

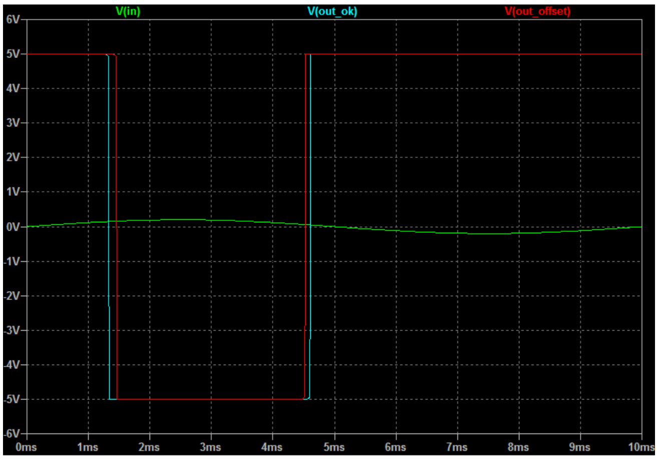


Fig.9. Simulation of effect of input offset voltage on a comparator

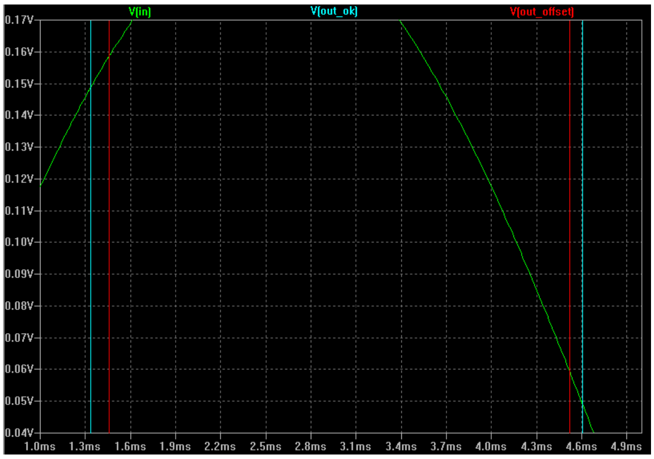


Fig.10. Zoom in on Fig.9 to show switching voltages

Symmetry

The closer the reference voltage is to the middle of the two comparator output voltages (the positive and negative supply voltages in the dual supply circuit in Fig.8) the closer we will be towards having the hysteresis evenly balanced around the reference. The circuit ***p_j*** is using has a single supply, so the two output voltages which set the two hysteresis thresholds are 0V and 5V (or more specifically the op amp's saturation output voltages, which may not be exactly equal to the supplies). The reference voltage is relatively low (under 1V on a 5V supply) so we would not expect the hysteresis to be very symmetrical with respect to V_{REF}.

Again, we can illustrate the issue with a couple of simulations. First, using the setup in Fig.11 we compare single (using U2) and dual (using U1) supply comparator circuits with the same reference voltage of 720mV. Both are driven by a positive offset sinewave (so that the input voltage stays at or above zero). The feedback resistors are different, but chosen to give about the same amount of hysteresis.

The results are shown in Fig.12, in which reference voltage V(ref) is shown as the green trace. The other two traces (red and cyan) are the V_{comp} voltages for the two comparators – not input or output voltages. These are the voltages at the op amps' positive inputs, which determine the input voltage at which the comparator switches. In both waveforms we see a voltage step as the comparator switches – this is the change of threshold from V_{TL} to V_{TH}, so the step size is the hysteresis.

The results in Fig.12 show that the hysteresis for the dual supply circuit (cyan trace, V(comp_dual)) is reasonably symmetrical with respect to the reference – the green V(ref) trace is about half way between the two levels of V(comp_dual). For the single supply circuit (red trace, V(comp_single)) the reference is much closer to the lower threshold level than the higher one.

The problem here is not a performance issue with the op amp or other circuit elements. The hysteresis is set by a combination of the reference voltage, the two op amp output voltages and resistor values. The two switching points can be whatever the designer wants, but because both V_{REF} and the two output voltages are scaled by the potential-divider effect, and the two output voltages are not necessarily equal and opposite, the two voltages are not symmetrical about V_{REF}.

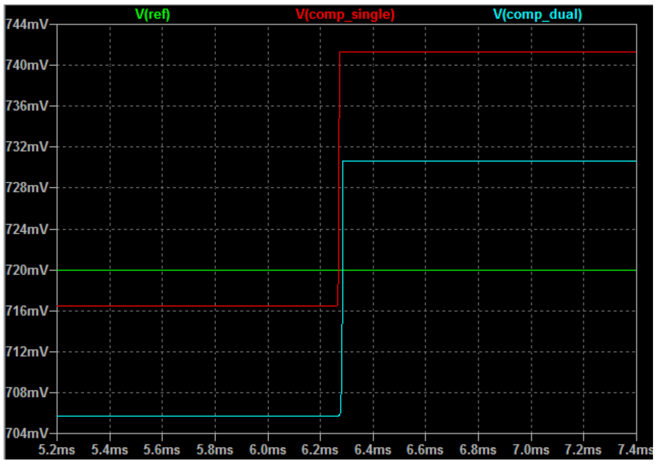


Fig.12. Simulation result from the setup in Fig.11. The green trace, V(ref) is the reference and the other two traces show the switching voltages of the two comparators

Potential divider

V_{REF} can be obtained from a potential divider connected to the supplies, but this does not behave like the ideal voltage sources which we have used for V_{REF} in the simulations so far – the circuit may load the potential divider unless it uses much smaller resistor values than those to set the hysteresis. An alternative approach is to let the comparator output directly interact with the potential divider to set the hysteresis; we no longer need R1 from the circuit in Fig.3. This version of the comparator is shown in Fig.13.

If we run the circuit in Fig.13 from a single supply, with supply voltage

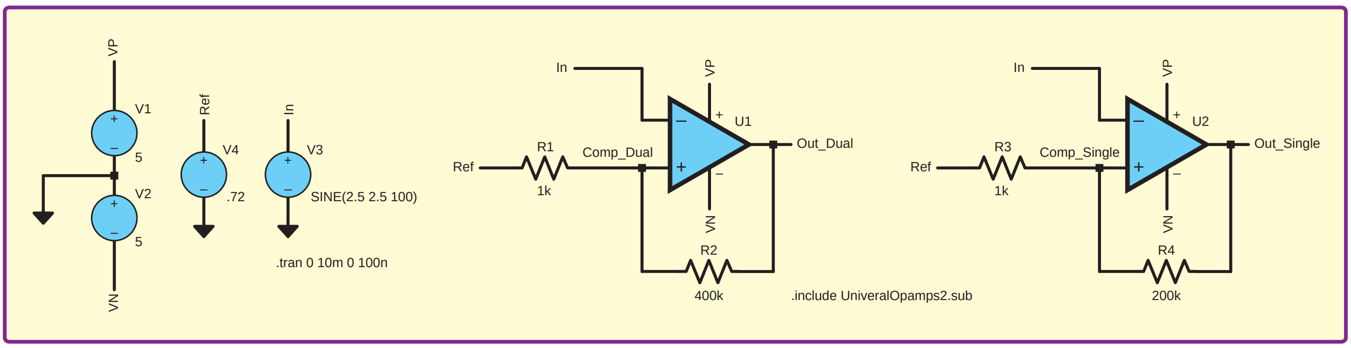


Fig.11. Simulation to compare single and dual supply comparators with the same reference and input signal

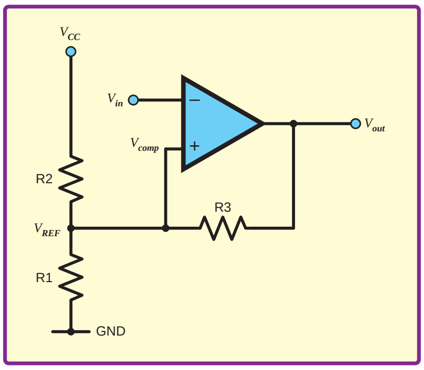


Fig.13. A potential divider can be used for the reference voltage to which hysteresis is applied. Single supply circuit shown

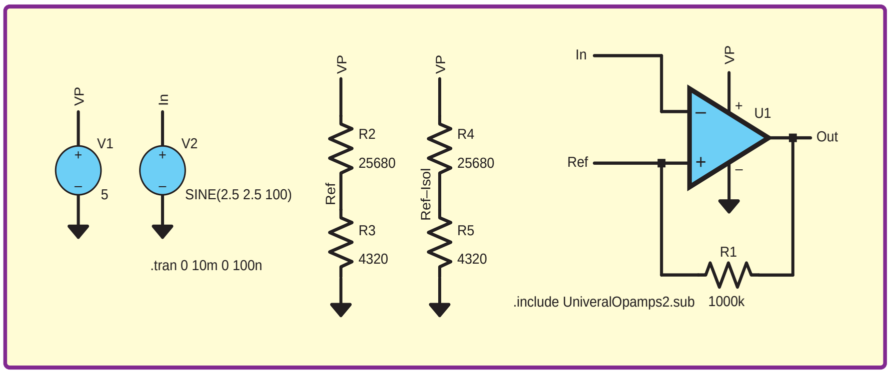


Fig.14. Simulation setup for the circuit in Fig.13

V_{CC}, and assume the op amp's output is perfectly 'rail to rail' (it outputs either 0V or V_{CC}) then:

$$V_{TH} = \left(\frac{R_1}{R_1 + R_{P23}} \right) V_{CC}$$

$$V_{TL} = \left(\frac{R_{P13}}{R_{P13} + R_{P2}} \right) V_{CC}$$

where R_{P13} is the parallel combination of *R1* and *R3* and R_{P23} is the parallel combination of R_2 and R_3. If the output is not perfectly rail to rail, then we need more complex equations, which include the two op amp output voltages as well as V_{CC}.

Resistor values

The switching points for the circuit in Fig.13 are not necessarily symmetrical around the open circuit voltage of the R_1, R_2 potential divider. For example, if we have V_{CC} = 5V and choose R_1 = 4.32kΩ and R_2 = 25.68kΩ we get a potential divider voltage of 720mV, with nothing else connected to it. If we have a feedback resistor R_3 = 1MΩ then the threshold voltages are V_{TL} = 0.717V and V_{TH}, = 0.736V, which is similar to the voltages quoted by ***p_j*** (we do not have full details of his circuit). We have a total hysteresis of about 20mV, which would be ±10mV about our 'set point' if the switching was symmetrical, but it is obviously not symmetrical as 717mV is much closer to 720mV than 736mV.

A simulation setup for this circuit is shown in Fig.14. The results of the simulation are shown in Fig.15, where we have zoomed in on one of the switching points. We have two equal potential dividers, one of them is not connected anywhere else (the Ref_Isol signal) and enables us to see the voltage of the unloaded divider (cyan trace). Here we can see that the two reference levels are close to the values just calculated, and as calculated, they are not symmetrical around the 720mV open circuit voltage of the divider.

If we want the circuit to switch at 720 ±10mV we can change the resistor values. For example with R_1 = 4.28kΩ, R_2 = 25.72kΩ (so the total is still 30kΩ) and R_3 = 900kΩ we get closer to 720±10mV. A simulation of this is shown in Fig.16. The V(ref_isol) potential divider (R_4, R_5) was left at 720mV for comparison purposes. The R_1, R_2 potential divider has an open circuit output voltage of about 713mV.

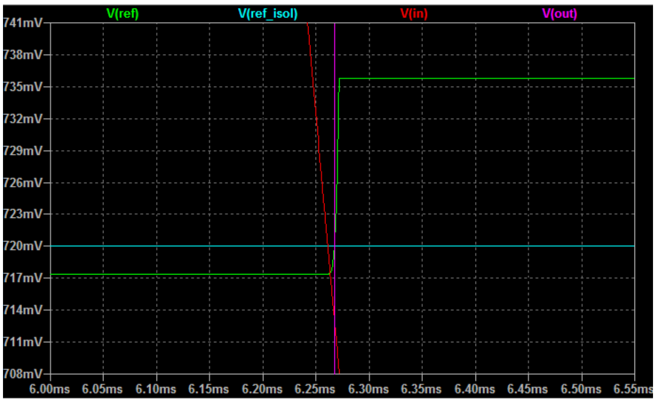


Fig.15. Simulation results for the circuit in Fig.14

Fig.16. Simulation results for the circuit in Fig.14 with new resistor values chosen to give more symmetrical switching around 720mV

Max's Cool Beans

By Max The Magnificent

Tick-tock, tick-?-

In recent *Cool Beans* columns I've mentioned my *Vetinari Clock*, which features four antique analogue meters: the large 'Hours' meter, two medium 'Minutes' and 'Seconds' meters, and a small 'Tick-Tock' meter that will swing back-and-forth like a miniature metronome. The clock also boasts an antique vacuum tube lit from beneath with tri-colored LEDs (here's a short video of the current state of play **http://bit.ly/195JKWR**).

The clock is named after Lord Havelock Vetinari from the *Discworld* books by Terry Pratchett (RIP). The clock in Lord Vetinari's waiting room keeps accurate time overall, but it sometimes ticks and tocks out of sync: 'tick-tock, tick-tock… tick-tock-tick… tock…' For anyone awaiting an audience, this is somewhat discombobulating, which is – of course – the whole point.

Thus, I need some way to make the 'tick-tock' sound, both for a regular 'tick-tock' mode and an irregular Vetinari mode. But I want more sounds than this. In fact, I'm thinking of having a variety of different modes, each with its own sound effects. In one mode, for example, I'd like to have sounds of clockwork and mechanical movements taking place in the background. Another mode might offer a mixture of hydraulic and pneumatic sounds, like a 'drip-drop' version of 'tick-tock', accompanied by occasional puffs of compressed air and suchlike.

Vetinari chimes

I personally hate the Westminster Chimes effect, unless we're talking about the real thing that originated at the church of St Mary the Great, Cambridge, and that is featured at the Houses of Parliament ('Big Ben'). However, I have contemplated implementing a Vetinari equivalent that misses off the terminal 'dong' (ie, 'Ding-Dong Ding-Dong; Ding-Dong Ding-?-').

But wait, there's more! In a regular analogue clock, as the minute hand circumnavigates the face of the clock, the hour hand slowly migrates from one hour to the next. This is one reason why young kids find it difficult to read the time, because they have to work out which hour we're talking about when we're part way between hours. I could replicate this on my 'Hours' meter, but I think I'd rather leave its needle pointing at the current hour until the 'Minute' meter reaches '60', at which time the 'Hours' meter can move to point to the next hour and the 'Minutes' meter will return to its starting position.

But it would be boring if all this happened without something to mark the event. Let's suppose that as we approach the end of the hour – say 10 seconds to go – the 'Hours' meter's needle starts to quiver and we hear a 'graunching' type of sound; then, on the hour, the 'Hours' needle springs to the next position accompanied by a 'releasing' type 'twang' effect.

Prototype jig for the **Inamorata Prognostication Engine**

Food for the ears

Actually, I want to deploy sound effects in quite a few of my projects. Take my *Inamorata Prognostication Engine*, for example. This is festooned with push-buttons and toggle switches and knobs. When I flick a toggle switch, I don't want to hear an embarrassing little 'click'. No! I want my ears to be tantalised with a rotund, robust, highly satisfying 'Ker-thunk!' What? You think that's it? Good grief. Is that how little you know me? *Au contrair*, I want the clicking of each switch to be followed by some audible accompaniment – food for the ears, as it were – such as the sound of balls rolling down shoots and complicated mechanical mechanisms performing strange and mysterious actions.

Ask not for whom the bell squeaks

Most of my projects are powered by Arduinos. Considering the humongous Arduino ecosystem that's out there, I fully expected to be able to find an appropriate sound-effects shield without much effort. Sad to relate, this is not the case. There are a lot of sound shields out there – like the Wave Shield from Adafruit (**http://bit.ly/1ykJRnI**) or the GinSing Synthesizer Shield (**http://bit.ly/1F2hNKJ**) – but none do what I want them to do.

I don't want to use synthesised sounds – I want to use small snippets of real-world recordings. The problem is that existing MP3-type cards tend to be geared up to playing a single audio stream, but I want to add layers of sound and merge different streams together on the fly. Since nothing suitable is readily available, I'm working on building such a shield with some chums. I will report further in a future column. In the meantime, if you have any ideas relating to interesting sounds for my projects, please feel free to email me at: **max@CliveMaxfield.com**. Until next time, have a good one!

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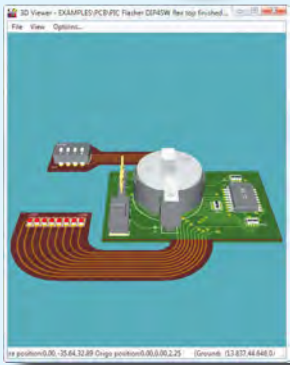

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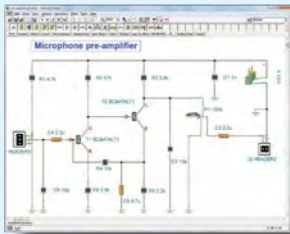

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The series is aimed at those using PIC microcontrollers for the first time. Each part of the series includes breadboard layouts to aid understanding and a simple programmer project is provided.

Also included are 29 *PIC N' Mix* articles, also republished from *EPE*. These provide a host of practical programming and interfacing information, mainly for those that have already got to grips with using PIC microcontrollers. An extra four part beginners guide to using the C programing language for PIC microcontrollers is also included.

The CD-ROM also contains all of the software for the *Teach-In 2* series and *PIC N' Mix* articles, plus a range of items from Microchip – the manufacturers of the PIC microcontrollers. The material has been compiled by Wimborne Publishing Ltd. with the assistance of Microchip Technology Inc.

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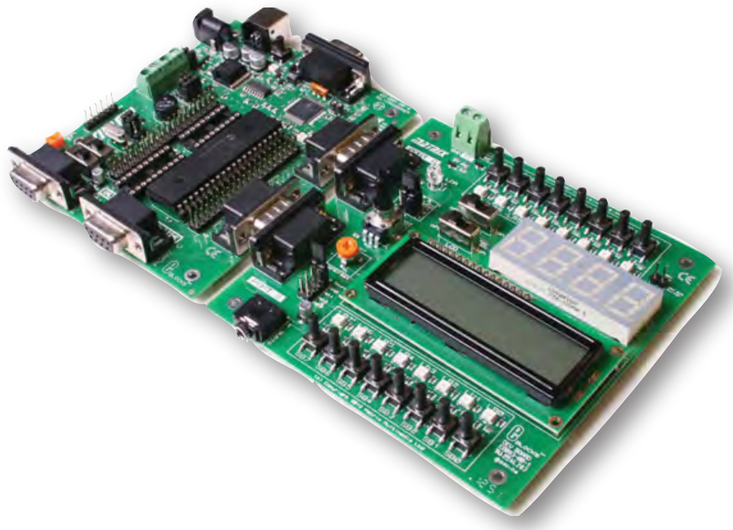

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(Formerly PICtutor)

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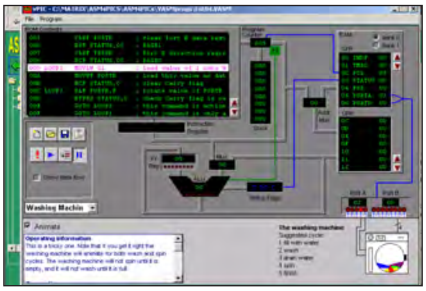

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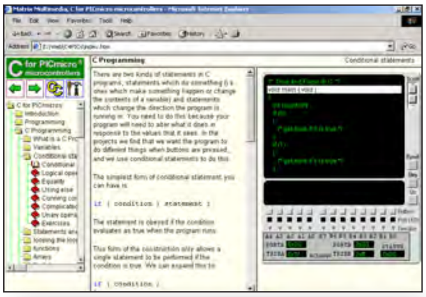

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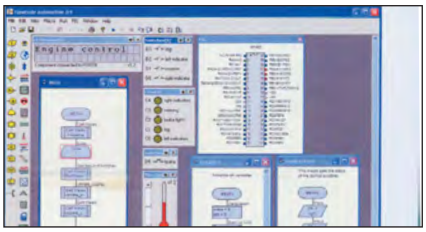

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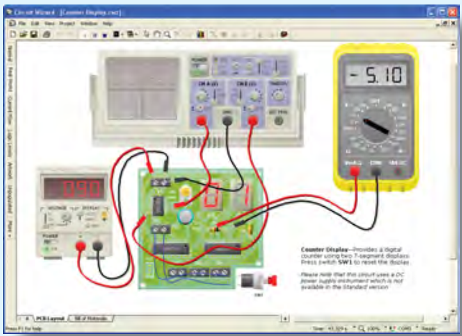

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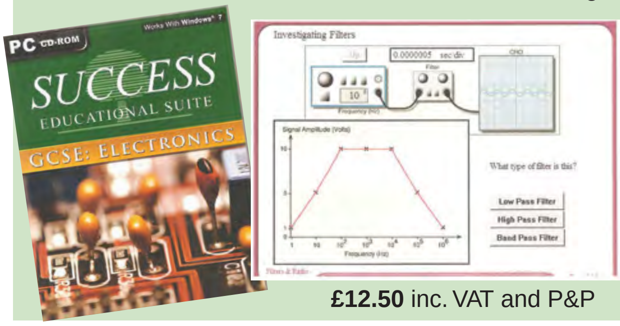

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PIC n' Mix

Mike Hibbett

Our periodic column for PIC programming enlightenment

Faster LCD interface

We reported in April's *PIC n' Mix* how one of our readers found a large LCD display that worked with the LPLC board, but that he had problems with the speed of the display update. The display provided 320 × 240 pixels in full colour, which meant writing to 230,000 pixels. That's a lot of data to write, especially when you are creating a project like an oscilloscope that must update the display several times a second. Although there are ways to minimise the amount of data that you change on each update (writing only the pixels that change, for example) the larger display highlighted an issue with our LCD driver software.

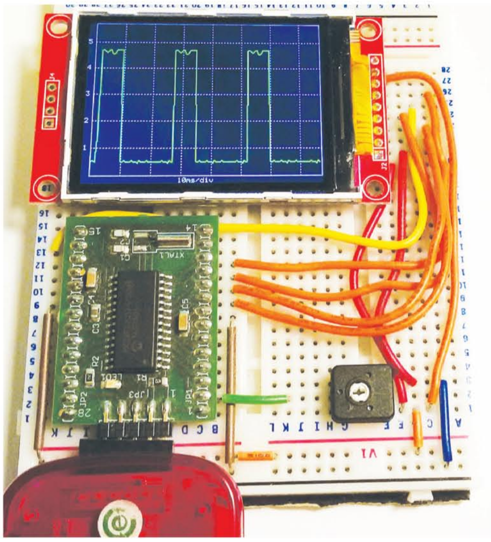

Fig.1. Simulated oscilloscope display

So we set up a test using the existing LCD driver on the LPLC board to see just how long it takes to write to the display. After power up, the display is cleared, then a fake oscilloscope display is drawn, as shown in Fig.1. The whole display update sequence was timed: 3.2 seconds. Let's see if we can improve on that.

Bit-bashing

The first inefficiency to deal with is the serial connection to the LCD, which is based on the SPI bus protocol. At the moment, we are using a 'bit-bashed' interface. This refers to how we control the signals that go between the microcontroller and the LCD, the serial clock and data signals. These signals can be seen on the schematic and timing diagram in Fig.2 and Fig.3 as the four wires CSx, D/Cx, SDI and SCL. A reset signal is also provided, to allow the processor to reset the LCD's internal state.

The offending code is limited to a single function:

```
void LCD_Writ_Bus(unsigned char da)

{

    LCD_SDA=(da & 0x80) ? 1 : 0; LCD_SCK = 0; LCD_SCK = 1;

    LCD_SDA=(da & 0x40) ? 1 : 0; LCD_SCK = 0; LCD_SCK = 1;

    LCD_SDA=(da & 0x20) ? 1 : 0; LCD_SCK = 0; LCD_SCK = 1;

    LCD_SDA=(da & 0x10) ? 1 : 0; LCD_SCK = 0; LCD_SCK = 1;

    LCD_SDA=(da & 0x08) ? 1 : 0; LCD_SCK = 0; LCD_SCK = 1;

    LCD_SDA=(da & 0x04) ? 1 : 0; LCD_SCK = 0; LCD_SCK = 1;

    LCD_SDA=(da & 0x02) ? 1 : 0; LCD_SCK = 0; LCD_SCK = 1;

    LCD_SDA=(da & 0x01) ? 1 : 0; LCD_SCK = 0; LCD_SCK = 1;

}
```

This code is fairly standard for bit-bashing a serial protocol like SPI. The same interface could have been achieved with a loop, but this technique is slightly faster, at the expense of a little extra code space. Even this code, however, is inefficient. Each of the eight lines of code actually contain three separate statements, so this is really 24 lines of C. And no matter how efficient the C compiler is, this will result in 30 to 60 machine code instructions at a minimum, possibly higher. Although our processor is fast, churning through twelve million instructions per second, the display requires many hundreds of thousands

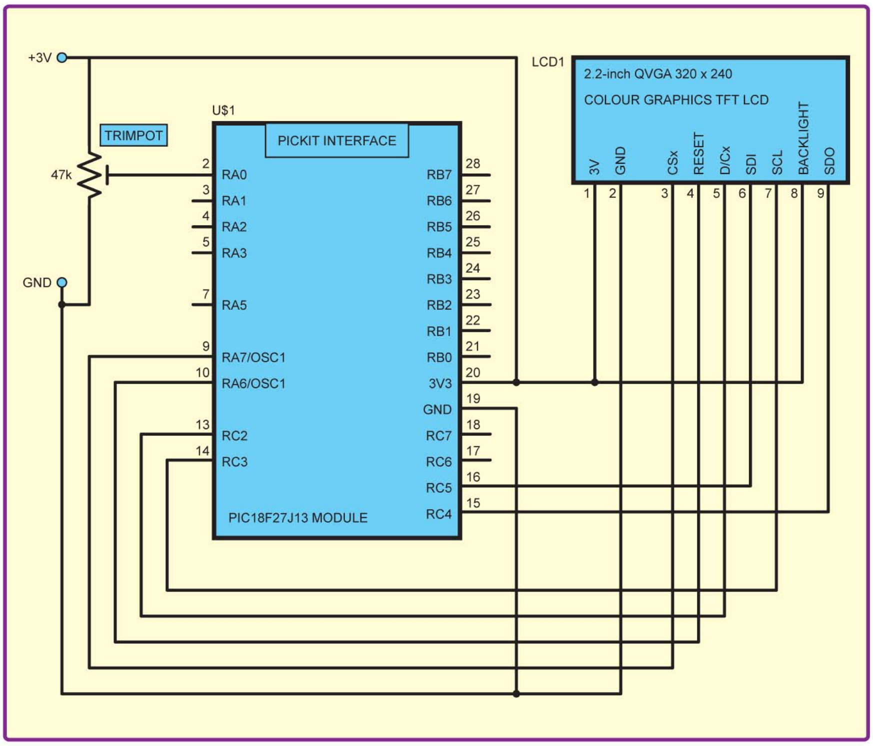

Fig.2. This month's circuit

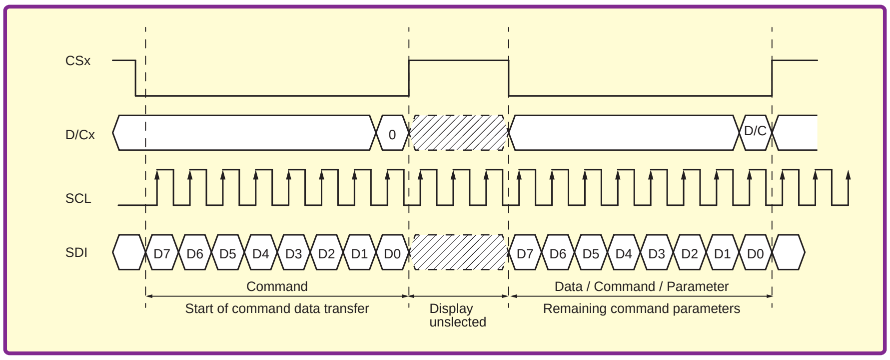

Fig.3. LCD control signal timing diagram

of calls to this code to fill it with an image. We need to do better.

This is why most PIC18F processors have an SPI peripheral built into the chip. A peripheral is a small section of the IC that is dedicated to a particular function, 'offloading' the work from the CPU to this dedicated electronic circuit. The benefits of this are significant, and varied – as a dedicated circuit it runs much quicker, consumes less power and even saves some code space. The code to use it is quite simple; two lines of code to initialise the SPI peripheral itself (selecting clock speed, 12MHz, and the edge of the clock on which the data will change) and then the code to send a data byte – just three lines of code, two of which are there only to stop us overfilling the SPI transmit buffer.

Here is the new, additional initialisation code that goes into the `Lcd_Init()` function:

```
// Configure the SPI peripheral.
    // 12MHz SPI clock
    SSP1STAT = 0x40;
    SSP1CON1 = 0x20;
```

The `LCD_Writ_Bus` function becomes shorter:

```
void LCD_Writ_Bus(unsigned char da)
{
    SSP1BUF;                  // Clear the BF flag
    SSP1BUF = da;             // write the data
    while (!SSP1STATbits.BF); // Wait for the tx to complete
}
```

The LCD initialisation and image display now takes only one second, more than three times faster. That's more like it!

There is a minor caveat to this speed improvement. Our interface between the microcontroller and the display is now running at 12MHz, and at these kind of speeds the interconnect between the two can become a bottleneck. If your wires are long the extra capacitance may degrade the signals, and cause the data signals to be received incorrectly. Our test with a breadboard and reasonably short hook-up wires (about four inches) worked, but twice that length and you may see problems, such as corrupted pixels or even complete failure to start the display. If that happens, reduce the length of your wires, or slow down the interface. You can halve the data rate and still see a performance improvement over the original code by changing the following line in the LCD driver code:

```
SSP1CON1 = 0x20;
```

to

```
SSP1CON1 = 0x2A;
```

Naturally, if you are building the circuit on a PCB or Veroboard then the interconnect will be much shorter and you can expect flawless operation at 12MHz, the maximum rate we can run the interface with our processor.

Can we improve on this further? We can, but by exactly how much will depend on each individual application. For our example application, an oscilloscope, we do not need to redraw the entire screen each time the trace changes; instead, our algorithm could be something like this:

```
loop:
  Draw the grid
  Draw the signal trace
  Acquire the new trace
  Draw the old trace in the background colour
  goto loop
```

Running some tests with the new LCD driver and a simulation of the above algorithm showed that we can achieve this 'loop' update in just 0.25s. Four times a second, quite reasonable for an oscilloscope display update. Excellent! So long as the time required to 'acquire the new trace' can be kept to a minimum, we should be able to create a workable oscilloscope. Acquiring the trace means reading samples through the ADC peripheral, which we will look at next month.

Even the above algorithm can be improved. Why do we waste time redrawing the grid? Because the trace crosses the grid lines, and when we erase the old trace we erase the points of the grid where the trace crosses it. You can see these grid lines in the example display in Fig.1. The next improvement will be to only redraw the individual points on the grid that get erased. Making that change will probably double the display update rate. We will save *that* improvement for next month.

Why do we even use bit-bashing if it is so inefficient? Not all micro-controllers have an SPI module, so in those cases there is no other option. If you don't have access to a digital storage oscilloscope it can also be difficult to debug an SPI peripheral – you write a few registers and cross your fingers that the code does what you expect. With bit-bashing, you can see in the code exactly what the signals will do, and you can single-step each line of code and measure the serial clock and data pins with a simple multimeter. Once you are comfortable that the series of bytes that you transmit work, and your external device communicates, then you can look at optimising the code. Then, if the device does not work, you know it is just the configuration of the SPI peripheral that is at fault, and not a million other possibilities. We've spent hours debugging a display interface, only to find we had forgotten to adjust the contrast!

PICKIT Programming interface

A common feature of every microcontroller-based design these days involves some form of programming interface. Chip vendors have their own interface (serial, JTAG, custom) and Microchip is no different. The days of having code held in an external memory device that needs to be 'popped' from the board and placed into a programmer are a distant memory (if you'll excuse the pun) and almost every design, in particular hobbyist designs, start with the need to place some kind of programming header on the board.

While some may grumble at Microchip's custom PICKIT or ICD interface, it's a simple enough six pin 0.1-inch SIL header and like many hobbyists we have a drawer full of them in the workshop. There are no additional components

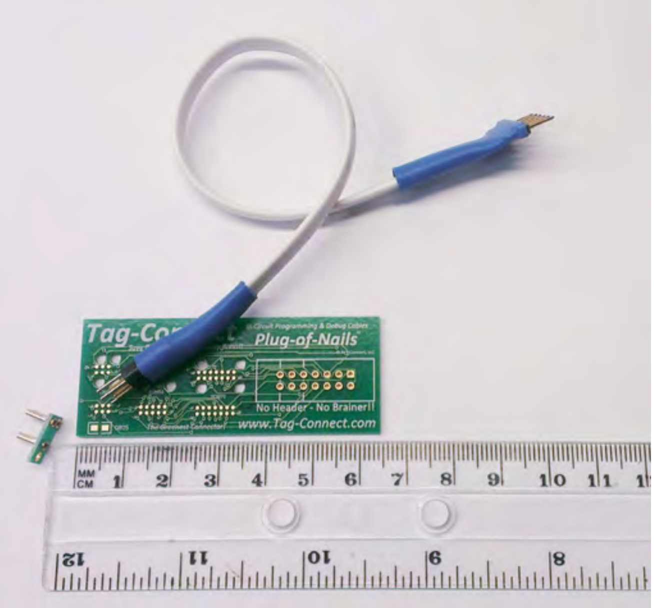

Fig.4. Tag Connect's simple programming interface

header and like many hobbyists we have a drawer full of them in the workshop. There are no additional components required, but it does take board space and require six holes to be drilled. If your interests lie in small board designs (such as wearable computing devices or tiny IoT sensors) then this additional board space can be an issue. Also, if you are a small company building products that have to be programmed by hand, the effort involved – fitting a SIL header, plugging in, programming, unplugging – can be a nuisance.

Fortunately there is an interesting commercial solution to this problem – a specialised programming connector that requires no mating header on your board, just three holes and six tiny pads on your board. Created by Tag Connect in the USA, the system consists of a small adaptor cable, shown in Fig.4. One end is a standard six-pin 0.1-inch SIL header to connect to your PICKIT, the other end is their custom header. It consists of three locating pins and six tiny spring-loaded pins. You simply push the header through your PCB, hold it in place and click 'program'.

For debugging, where the PICKIT needs to be connected for long periods of time, Tag Connect supply a very simple latching PCB that will hold the locating pins in place. The board space required for this is just 9mm × 5mm. Tag Connect supply symbols for a number of different PCB CAD programs, so it is easy to add to your PCB designs. The cable is a bit pricey at 34USD, but like the PICKIT, it's a tool that should last for many years, and if you are in the business of building many boards – or really need that extra board space – this could be an ideal solution. We wish we'd had it for the LPLC TOO board design! For further details, visit: **www.tag-connect.com**

Maker Faire

While Maker Faire UK is just behind us, Ireland's is yet to come – the Dublin Maker Faire is set for 25 July. If you've not come across one of these events yet, they are an opportunity for hobbyists to demonstrate their electronics, mechanical, craft or artist skills to the general public. They are fun community events that offer inspirational, amusing and sometimes downright daft ideas. We attended our first Maker Faire last year and are now hooked. The venue hasn't been secured yet for the Dublin event, but for those of you who might be able to make it we will publish the venue once known. With luck, the weather will be kind to us again this year. Last year, it only rained a bit, rather than a lot. Sunshine would be just too much to ask for!

Next month

Next month, we look at reading an ADC port in real time, acquiring data that we display on the LCD – in short, a simple oscilloscope. We will keep the hardware interface simple, but hopefully with the board and the software created so far we should be able to see real analogue signals on the LCD.

Not all of Mike's technology tinkering and discussion makes it to print. You can follow the rest of it on Twitter at ***@MikeHibbett****, and from his blog at* ***mjhdesigns.com***

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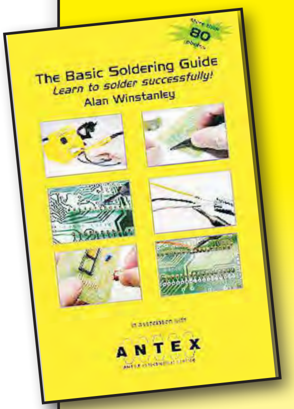

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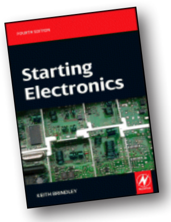

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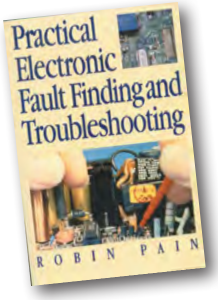

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All prices include VAT and postage and packing. Add £2 per board for airmail outside of Europe. Remittances should be sent to **The PCB Service, Everyday Practical Electronics, Wimborne Publishing Ltd., 113 Lynwood Drive, Merley, Wimborne, Dorset BH21 1UU. Tel: 01202 880299; Fax 01202 843233; Email: orders@epemag.wimborne.co.uk. On-line Shop: www.epemag.com**. Cheques should be crossed and made payable to *Everyday Practical Electronics* **(Payment in £ sterling only)**.

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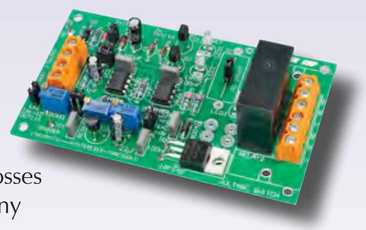

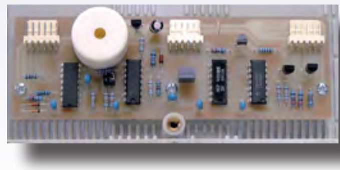

Threshold Voltage Switch

Here's a very handy, simple but versatile device to switch a relay when an input voltage crosses a preset threshold. This project takes the output of an analogue sensor, battery voltage or any other varying voltage and switches power to a buzzer, fan, warning light or similar when a preset threshold voltage is reached. It can also be used to prevent a lead-acid battery from being over-charged.

L-o-o-o-n-g Gating Times for the 12-Digit High-Resolution Counter

This add-on PCB module enables higher resolution measurements with the *12-Digit Frequency/Period Counter* described in the January to March 2014 issues of *EPE*. It adds an additional decade divider for the external timebase input to allow measurements using a gating time of 10,000 seconds (nearly three hours) and includes front-panel signal LEDs for gating indication.

Touch-Screen Digital Audio Recorder – Part 2

Next month, in Part 2, we will give the assembly details, provide some performance graphs and describe how this exciting and fun-to-build *Touch-Screen Digital Audio Recorder* is used.

Teach-In 2015 – Part 6

July's *Teach-In 2015* maintains the momentum of learning linear design as we demonstrate our favourite software to check the operation of our tone control circuit, and also how we measured the final prototype's performance. We will also examine power and power measurement, and look at two useful circuit building blocks – constant current and constant voltage sources.

JULY '15 ISSUE ON SALE 4 JUNE 2015

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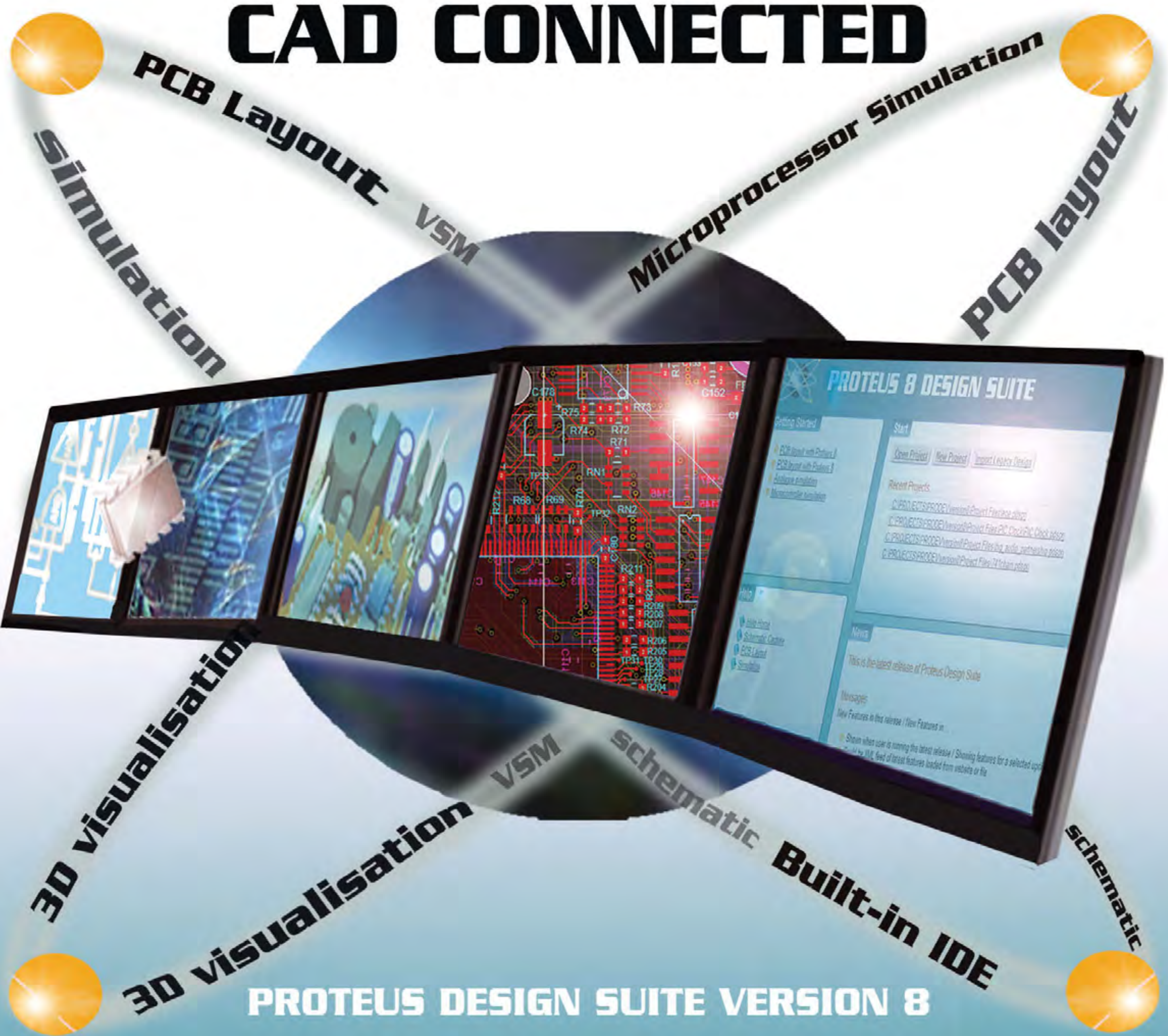

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